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CHAPTER VI

RADAR ANTENNAS

Section 6-1. Antenna Radiation Patterns

Antennas employed in radio stations are classified by type of transmission and reception as directional or non-directional.

Non-directional antennas emit electromagnetic energy uniformly in all directions and receive it regardless of the direction from which the incoming radio waves arrive.

On the other hand, the magnitude of the e.m.f. induced in a directional receiving antenna in a given position and constant intensity of the swept field is highly dependent upon the direction of the received wave, i.e., on the orientation of the receiving antenna relative to the transmitting station. A directional receiving antenna amplifies the reception of waves arriving within a narrow directional angle, and weakens reception of waves coming from other directions.

Likewise, when directional transmitting antenna is in a given position, the intensity and strength of the electromagnetic field which it emits depends, at points equidistant from the antenna, on the direction. A directional transmitting antenna concentrates the emission into a narrow beam in a single direction and weakens the radiation in other directions.

The application of the so-called reciprocity theorem is very helpful in antenna engineering. This theorem states that the nature of the directional activity of an

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antenna is absolutely identical, whether it is used for transmission or reception.

The directivity of an antenna is conveniently depicted by a radiation pattern.

The radiation pattern or characteristic of an antenna is the ratio of the emitted antenna field to the direction toward points of reception at identical distances from the antenna.

A distinction must be made between the radiation pattern for the field intensity E (the e.m.f., \mathcal{E} , induced in a receiving antenna) and that for the radiated power P (picked up by the receiving antenna).

We know from the theory of the electromagnetic field that the power of a field is proportional to the square of its intensity. Consequently, as the direction changes, the power of a field changes very much less than its intensity. The pattern of an antenna is always much more distinct relative to the field power than to the intensity.

The directivity of an antenna is complex in nature and manifests itself differently in the horizontal and vertical planes. Therefore, its pattern in the horizontal plane (along the azimuth ϕ) is different from that in the vertical plane (along the elevation θ).

For the purposes of generalization from these concepts, we shall speak of antenna radiation patterns at a given angle α , (instead of ϕ or θ) relative to a magnitude A , of interest to us (instead of the intensity E , or power P).

A pattern of this type may be expressed analytically both in absolute scale, in the form of the equation $A=F(\alpha)$, and in relative scale, in the form of the equation $\frac{A}{A_{\max}}=F_1(\alpha)$. Diagrams of radiation patterns plotted on absolute or relative scales are geometrically similar, since they differ only in scale, and merge into an identical curve when the scales are chosen accordingly. The term radiation pattern is often applied to these diagrams.

Figure 4-1 presents such a pattern in a polar system of coordinates.

In any given radiation pattern, the magnitude of A or of $\frac{A}{A_{\max}}$ may be deter-

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mined for the given angle α .

In plotting a radiation pattern, its center O is chosen, and the line from which the angles α are to be measured off, is indicated. In order to determine the magnitude A or $\frac{A}{A_{\max}}$ by means of the antenna pattern in a direction in which we are interested, a radius-vector is drawn from the center of the diagram at the desired angle α . The length of the vector determines A or $\frac{A}{A_{\max}}$ for the angle α of interest

a)

b)

A_{\max}

Line
from
which
angles
are
measured

Fig.6-1 - Antenna Radiation Pattern in Polar Coordinates

a) In absolute scale; b) In relative scale, with beam width γ .

to us. The ends of all the radius-vectors are defined by the outline of the pattern, and they begin at the starting point of the pattern.

Note that the definition of the pattern is very often determined by the so-called beam width, by which is meant the angle determined by those points in the pattern where the radiated power is equal to one-half of the maximum power in the pattern (Fig.6-1).

The radiation pattern of a sending antenna may be determined experimentally as follows. First, points are selected around the station antenna C, at uniform distances therefrom (say, one km.) in accordance with the wanted angles α (Fig.6-2). A special indicator is used to measure the magnitude of A at all these points successively. The experimental data (α and A) are tabulated, and this compilation is used to plot the pattern on an absolute scale $A=F(\alpha)$. To plot the pattern in rela- STAT

tive scale $\frac{A}{A_{\max}}$, we find the maximum radius-vector A_{\max} on the prior pattern. The magnitude A in the Table for all angles α is divided by A_{\max} .

It should be noted that when a directional transmitting antenna is rotated, this motion proceeds in the direction of maximum emission. Thus, the rotation of a direc-

Indicator of
magnitude of A

Fig.6-2 - Method of Determining
Radiation Pattern

Fig.6-3 - Directional Antenna
Pattern

tional antenna is equivalent to the rotation of its radiation pattern in the opposite direction.

In non-directional antennas, the magnitude of A is not dependent upon the angle α . The radiation pattern of such an antenna is a circle (Fig.6-3).

When plotting the vertical pattern of an antenna, it is necessary to take into account the effect of the ground, consisting of the reflection by the earth's surface of radio waves radiated by the antenna toward the ground. Let us look at Fig.6-4a. At the point T_s , the direct antenna ray and the reflected ray are superimposed. The resultant field at the target will depend upon the difference in course of the direct ray AT_s and the reflected ray AOT_s .

The reflected field may be replaced approximately by a mirror image of the antenna A_{20} , constituting an identical antenna, located below ground at a depth equal to the height of the real antenna above ground (Fig.6-4a). The currents in the horizontal portions of the antenna and in its mirror reflection must be regarded as flow-

ing in the same direction, while the currents in the vertical portions of the antenna and its mirror reflection are to be regarded as flowing in opposite directions (Fig.6-4b). The ground itself is entirely disregarded.

In certain directions, the direct ray and the specular reflection ray are in

a) Direct ray b)

Reflected ray

c)

Fig.6-4 - Effect of Ground on Vertical Antenna Pattern

phase on arrival at the point of intersection, causing an approximate doubling of the gain of the resultant field. In certain other directions the direct ray and the specular reflection ray arrive at the point of intersection in counterphase, so that the resultant field is zero. Ground reflection gives the vertical pattern a multi-lobed appearance (Fig.6-4c), the number of lobes in the vertical diagram being equal, as shown by the antenna theory, to the number of half-waves represented by the elevation of the antenna above the ground. Thus, if the antenna is elevated above ground to a height equivalent to $3 \frac{\lambda}{2}$, where λ is the working wavelength of the station, the vertical pattern of the antenna will have three lobes.

The basic requirements for the antenna system of a radar may be formulated as follows:

a) it must provide a radiation pattern of the desired shape (and sharpness), assuring the required concentration of radiation in the direction of the target, the given accuracy of measurement of the angular coordinates of the target in direction finding, and the required bearing discrimination along the angular coordinates;

b) its design must be as simple as possible, mechanically strong, small and light-weight, handy in terms of rapid assembly and disassembly. These are the fundamental factors in determining the time required to mount the station for action from its transport set-up, and vice versa. The design of the antenna system must assure the required method and speed of sweep scanning by the electromagnetic ray, corresponding to the given type of radar.

Section 6-2. The Half-Wave Dipole

The basic component of most radar antennas is the half-wave dipole consisting of two identical metal tubes, usually silverplated, one-quarter the length of the broadcast wave. The total length of the half-wave vibrator is $\frac{\lambda}{2}$.

Insofar as distribution of the current along its length is concerned, the half-wave dipole is similar to a quarter-wave segment of a two-wire line open at the end. The site of the current loop and the voltage node is at the center of the dipole, while the voltage loop and the current node are at the ends (Fig.6-5a and b).

Electromagnetic oscillations around wires 1 and 2 of the quarter-wave segment are generated in antiphase to each other and compensate each other completely, so that radiation by this segment is quite insignificant. To do away with this compensation, wires 1 and 2 are rotated and placed in a straight line. A quarter-wave segment so rotated is converted to a half-wave dipole (Fig.6-5b). In both halves of the dipole the currents move in the same direction, an entire half-wave of current or voltage distribution being laid out along its entire length. The lines of forces of the electrical field of a half-wave dipole are directed primarily along its wires, while the lines of force of the magnetic field surrounding the wire in concentric

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circles, are normal to the wires.

The plane passing through the center of a half-wave dipole perpendicular to its wires is generally termed the equatorial plane.

The intensity of the electrical and magnetic fields of a dipole is dependent upon distance. At a given distance from the axis of the dipole, the electrical field

a)

b)

Fig.6-5 - Conversion of a Quarter-Wave Open-End Line (a)
to an Ordinary Half-Wave Dipole (b)

in all directions in the plane of the center-line is of identical intensity E_ϕ (Fig.6-6a). The same may be said of the intensity of the magnetic field H_ϕ , and the radiated power density, i.e., the Umov-Poynting vector modulus P_ϕ which is determined by the product of the effective values of E_ϕ and H_ϕ . Thus, a half-wave dipole radiates an electromagnetic field equally in all directions in the center-line plane perpendicular to the dipole axis. This means that in the center-line plane, the half-wave dipole has no directivity either in field intensity or radiated power. The center-line plane pattern of a half-wave dipole is a circle (Fig.6-7a).

Any plane passing through the dipole axis is called a meridional plane. The radiation of a half-wave dipole in a meridional plane is already clearly directive in nature, as it is dependent upon the angle θ between a line perpendicular to the dipole axis at its center and the direction which interests us (Fig.6-7b). In simplified form, this may be clarified by the following example (Fig.6-6b). The inten-

sity of the electrical field E_θ determining the radiated power in the direction of interest, is perpendicular to the Umov-Poynting vector \bar{P}_θ and is equal only to a projection of the vector \bar{E} at the given point in a direction perpendicular to that of interest to us.

The same may be said of the intensity of the magnetic field H_θ . Figure 6-6b

a) $\begin{matrix} E \\ H \end{matrix}$

b)

$\begin{matrix} E \\ H \end{matrix}$

Lines of magnetic force

Lines of electrical force

Lines of electrical force

Lines of magnetic force

Fig. 6-6 - The Field of a Half-Wave Dipole in the Equatorial (a)
and Meridional Planes (b)

makes it possible to establish the cosinusoidal relationship between the angle θ of the electrical and magnetic components of the field of a half-wave dipole radiating in the direction of interest to us:

$$\left. \begin{aligned} E_\theta &= E \cos \theta \\ H_\theta &= H \cos \theta \end{aligned} \right\} \quad (6-1)$$

The maximum intensity of the electrical and magnetic field is created in the direction perpendicular to the axis of the dipole ($\theta = 0$). On the other hand, in the direction of the dipole axis in which $\theta = 90^\circ$, the field is absent. The field

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intensity pattern of a half-wave dipole in the meridional plane resembles a figure eight and is shown in Fig.6-7b as the fine outline.

At the given point, the Umov-Poynting vector is proportional to the square of the intensity of the electrical or magnetic field at the same point. Therefore the radiation pattern of the radiated power of a half-wave dipole in the meridional plane (heavy line, Fig.6-7b) is sharper than the pattern discussed above, as it is subject to the square of the cosine:

$$P_{\theta} = P \cos^2 \theta \quad (6-2)$$

Be it noted that the considerations we have set forth lead to a clear concept of the directivity of the half-wave dipole. However, these considerations are not to be considered wholly rigid. In actuality, the electromagnetic field created by a half-wave dipole at the point of interest to us is the result of superimposition of fields induced by all of its components of infinitely short length, acting as elementary independent sources of radiation. In view of the inequality in the distribution

Fig.6-7 - Half-Wave Dipole in the
Equatorial (a) and Meridional (b)
Planes

of current along the length of the dipole, the elementary fields at the given point differ from each other in magnitude and also differ in phase due to the difference in distances from the various components at the given point. Therefore, the actual directivity of a half-wave dipole is somewhat less than would be indicated by a simple cosinusoidal equation in the meridional plane, and the pattern emerging is slight-

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ly flattened. On the same principle, a half-wave dipole reveals a barely noticeable directivity in the equatorial plane as well. However, for all practical purposes this may be left out of consideration to simplify the reasoning.

Now let us turn to the question of the resistance of a half-wave dipole to radiation. Resistance to radiation by an antenna R_{Σ} is that conditional active antenna resistance at which the current feeding the antenna I develops the power P_{Σ} radiated by the antenna in the form of radio waves.

According to this definition, the resistance to radiation is equal to the ratio of the total radiated antenna power to the square of the true magnitude of the current at the point where the antenna is fed by the generator oscillator:

$$R_{\Sigma} = \frac{P_{\Sigma}}{I^2} \quad (6-3)$$

As shown in the theory and confirmed in practice, the resistance to radiation by an isolated half-wave dipole is $R_{\Sigma} = 73.2$ ohms regardless of the length of the operating wave, so long as the dipole is one-half its length and regardless of the manner in which the dipole is positioned, provided that only a single dipole is being used. The resistance to radiation of the half-wave dipole determines the active component of its total input resistance loading the supply feeder.

A major shortcoming of the ordinary half-wave dipole is its low resistance to radiation and therefore its low resistance to the load on the supply feeder. In order to attain a high traveling-wave ratio in the supply feeder, its wave resistance should also be of the order of 73 ohms. Manufacture of a feeder of such low resistance to transmit power of the order of hundreds of kilowatts is quite difficult.

In order to obtain low $\rho = \sqrt{\frac{L_1}{C_1}}$, it is necessary to make sure that the linear inductance L_1 is low and that linear capacitance C_1 is high. Both demand the use of thick wires closely spaced, and this in turn limits the permissible working voltage on the feeder and the power it delivers (see Section 4-7).

The use of half-wave folded dipoles instead of the ordinary kind has two ob-
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a) to increase the radiated power with no change in the current in the feeder line, i.e., to increase the resistance to radiation;

b) to make it possible to use feeder lines of higher resistance without the need for resistance transformers.

The folded dipole may be regarded, in terms of longitudinal distribution of current, as a two-wire line $\frac{\lambda}{2}$ long, folded into a loop and shorted at the end (Fig.6-8). Due to the equality of the currents at the upper and lower edges of the

a)

b)

c)

Fig.6-8 - Folded Dipole

a) Shorted half-wave line; b) Same line, converted to a folded dipole;

c) Folded dipole as parallel connection of two elementary half-wave dipoles.

loop, the folded dipole consists of a system of two elementary half-wave dipoles connected in parallel and combined into one. A folded dipole is therefore sometimes called a double doublet or a "bivibrator".

The fields excited by the upper and lower elementary dipoles constituting the folded dipole are superimposed. The voltage of the resultant field of the folded dipole will at any point be twice the field intensity of the elementary dipole, i.e.,

it will be $2E_1$. Therefore the power density and the total radiated power of a folded dipole P_Σ are proportional to the square of the voltage $(2E_1)^2$ and will come to $4P_{\Sigma 1}$. i.e., four times the power radiated by an ordinary half-wave dipole. Given the same amount of current in the feeder, we arrive at the following expression for the radiation resistance of a folded dipole:

$$R_\Sigma = \frac{P_\Sigma}{I^2} = 4 \frac{P_{\Sigma 1}}{I_1^2} = 4R_{\Sigma 1} \quad (6-4)$$

This means that a folded dipole burdens the feeder with an active component of input impedance $R_{in} = 4 \times 73.2 \approx 300$ ohms, or four times that of an elementary half-wave dipole, while on supply by a high-resistance feeder ($\rho \approx 300$ ohms) it does not require that a go-and-return resistance transformer be cut in, as it performs that function itself. For this reason a folded dipole is also sometimes called a loop dipole.

The radiation pattern of a folded dipole does not differ in any way from that of an ordinary half-wave dipole, which we have already examined.

Section 6-3. The Functioning of Half-Wave Dipoles in the Antennas of Very-Short-Wave Systems

The major shortcoming of the single half-wave dipole is its relatively broad radiation pattern, inadequate for the precision of measurement of angular coordinates and the degree of concentration of radiation toward the target usually required. The power gain of a half-wave dipole is only 1.64. A half-wave dipole is therefore to be considered a very primitive antenna and renders the use of single-unit installations impractical.

The behavior of a half-wave dipole changes very much for the better in the presence of other dipoles arranged in proper fashion.

Let us examine a system consisting of two-co-linear half-wave dipoles and at their centers separated by the distance d . If the feed to the dipoles is cophasal,

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i.e., if the phase of the current in both is identical, the electromagnetic fields induced around them will also be in phase. In accordance with the equation for the electrical component of the field of a radiated radio wave, then at r distance from the dipole, the following instantaneous voltage will exist:

$$e(r) = E_m(r) \sin \left(\omega t - \frac{2\pi}{\lambda} r \right)$$

The phase of the oscillations lags in proportion to the distance from the source of radiation, while the amplitude $E_m(r)$ declines as the first power of the distance. It may be reckoned that at a sufficiently distant point in space the electromagnetic waves may be taken as being emitted from the antennas in parallel fashion, so that in the general case, the phase difference in the plane shown in Fig.6-9 will depend on the angle θ :

$$\Delta r = d \sin \theta \quad (6-5)$$

The difference in the path of the rays will cause, at the point in which we are interested, a difference in the phases of the fields radiated by the antennas:

$$\Delta \varphi = \frac{2\pi}{\lambda} \Delta r = \frac{2\pi}{\lambda} d \sin \theta \quad (6-6)$$

Therefore, at the point in which we are interested, the antenna fields form a geometric progression, the result of the superimposition depending upon the angle θ so that in different directions the figures for the same distance from the radiating system will be different.

The difference in the amplitudes of the fields may be ignored in view of the small size of Δr relative to the distance from r to the point of interest. In accordance with equ.(6-5) and (6-6), $\Delta r = 0$ and $\Delta \varphi = 0$ in the direction perpendicular to the axial line of the system ($\theta = 0$), so that the shaping of the fields is simplified to an arithmetical progression, and the resultant field will be twice as powerful as that of a single dipole. In any direction other than the perpendicular, some difference in path and phase difference appears. The resultant field will not be twice that of a unit antenna, but somewhat less, as the progression will be geo-

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metrical and not arithmetical. Thus, a system of two cophased antennas clearly produces a more elongated pattern than does one from a single half-wave antenna.

In certain directions, the difference in path Δr is $\frac{\lambda}{2}$, and the phase differ-

Fig.6-9 - Two Antennas in Colinear Array

ence is $\Delta\phi = \frac{2\pi}{\lambda} \cdot \frac{\lambda}{2} = \pi$.

In such directions the antenna fields reach the point of interest in antiphase, and superimposition results in complete compensation, so that the resultant field of the system is zero. The angles θ in these directions determine the angular boundaries of the lobes of the system's pattern. The width of the lobe will depend upon the distance d between the antenna centers, and upon the length of the working wave λ . In the direction of the axial line the field of the system will always be zero, since none of the antennas radiates in this direction.

Figure 6-10a shows the radiation pattern of a system of two cophased antennas at the distance $\frac{\lambda}{2}$ between centers. When a large number of antennas is used, the pattern becomes even more elongated.

Another diagram results when the phases of the feed to a two-antenna system are in opposition and the currents in the antennas each show a 180° displacement of phase relative to the other. The antenna fields will also be excited in antiphase. In this situation there will be added to the phase difference in any direction a constant compensating phase difference π , determined by the fact that the ant

oscillations are in opposing phase. Therefore, in the direction perpendicular to the axial line ($\theta = 0$), the resultant of the field will be not at its maximum, but equal to zero, since the total difference in phase will equal π . As previously, the field

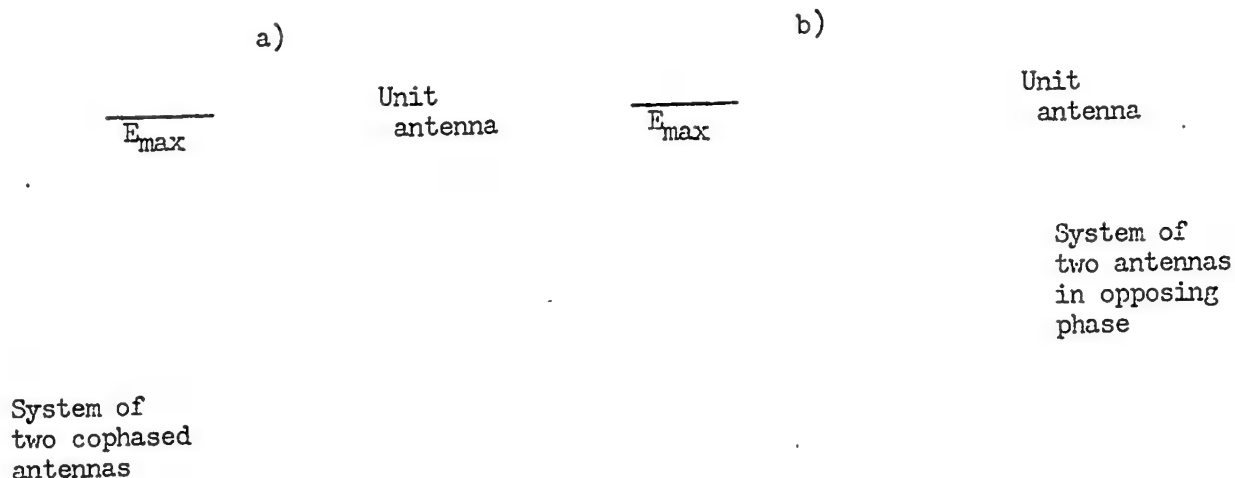


Fig. 6-10 - Two-Antenna Patterns

a) With cophasal feed; b) With feed in opposing phases

will equal zero in the direction of the axial line of the system as well, as none of the antennas radiates in that direction.

The maximum resultant field will be in those directions in which the difference in the path of the electromagnetic fields of the antennas compensates for the opposition of their phases as driven. When compared to the preceding case, the pattern of the system doubles symmetrically relative to the direction perpendicular to the axis. The angle of inclination of the lobes of the pattern from that direction, and the width of the lobes, depends upon the distance d between the antennas, and the

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wavelength λ .

Figure 6-10b shows the pattern when two antennas, the centers of which are $\frac{\lambda}{2}$ apart, are fed in opposite phase.

Section 6-4. Multi-Antenna Cophasal Arrays

Stations broadcasting in the very short wave band employ multi-unit cophased arrays, constituting systems of half-wave dipoles arranged in tiers. The dipoles in each tier are arranged in a single aerial line.

Figure 6-11a presents a cophased antenna consisting of four tiers of four dipoles each. The antenna is fed through a branched system connected to the transmitter. To assure a high traveling-wave ratio matching quarter-wave feeder transformers AT-1, AT-2 and AT-3 are inserted at the junction points of the feeder system.

It is necessary for the dipole feeds to be in phase in order to prevent the appearance of excess lobes in the pattern, as we saw them in the case of the feeding of a system of two dipoles in opposing phase. The traveling wave appearing in the feeder system at the ends of each half-wave feeder segment presents instantaneous phases differing by 180° . As shown by Fig. 6-11a, the length of each feeder segment between tiers is $\frac{\lambda}{2}$. Therefore, in order to assure cophasality, the terminals of every other tier are crossed on connection with the feeder.

As we have seen in the case of two cophased dipoles, the maximum radiation of a cophased antenna occurs in two diametrically opposite directions normal to the plane in which the dipoles are located.

An increase in the number of dipoles in each tier narrows the pattern in the horizontal plane, i.e., along the azimuth. The patterns of all tiers are identical and cophasal. In space they are arithmetically superposed, amplifying the resultant field of the system in the horizontal plane by an identical factor in each direction, this factor being proportional to the number of tiers. Consequently, an increase in the number of tiers does not affect the shape of the pattern in the horizontal plane.

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On the other hand, an increase in the number of tiers compresses the pattern in the vertical plane, i.e., along the tilt, due to the existence in this plane of a

a)

to transmitter

b)

Induction bridges

Fig.6-11 - Multi-Unit Cophased Antenna:

a) Active array; b) Parasitic reflector

difference in the paths of the rays of the dipoles in different tiers of a vertical series. Just as the number of tiers does not affect the shape of the pattern in the horizontal plane, an increase in the number of vertical series, i. e., of the number of dipoles in each tier, does not sharpen the pattern in the vertical plane.

Figure 6-11a shows the pattern in the horizontal plane, while Fig.6-12 illus-

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trates a multi-dipole antenna. The width of the main horizontal lobe in the pattern is about 15° . The main lobe in the vertical pattern is of approximately the same width.

When antennas with large numbers of dipoles are used, an even sharper pattern may be attained, but in this case the antennas themselves become overlarge both in height and width.

It is obvious that antennas having this type of horizontal pattern cannot be used in radar. The radiation of the antenna is concentrated irrationally in two dia-

Fig.6-12 - Directivity Pattern of the Exciter of a Multi-Unit
Cophased Array Along the Azimuth φ (in Horizontal Plane)

metrically opposite directions. In reception with such an antenna it is probable that there will be an error of 180° in azimuthal direction finding. Therefore, elimination of the second major lobe in the horizontal pattern is most important in designing the antenna system. The solution employed in radar has been developed from the work done at the Nizhniy Novogorod Radio Laboratory by K.A.Bonch-Bruевич, V.V.Tatarinov and A.A.Pistol'kors, who developed the world's first directional short-wave antennas.

Parallel to the plane of the cophased dipoles, which we term the exciter, and at a distance of $\frac{\lambda}{4}$, an identical system of dipoles is setup which is called the parasitic reflector (Fig.6-11b).

The reflector dipoles are not driven from the transmitter. The current they

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contain is induced by the electromagnetic field of the exciter. The reflector merely tunes into the working wavelength by means of induction bridges. The functioning of an antenna array with a reflector is readily understood with the aid of a vector diagram (Fig.6-13).

The magnetic field of the exciter A, designated as \vec{H}_{aa} in the vicinity of the

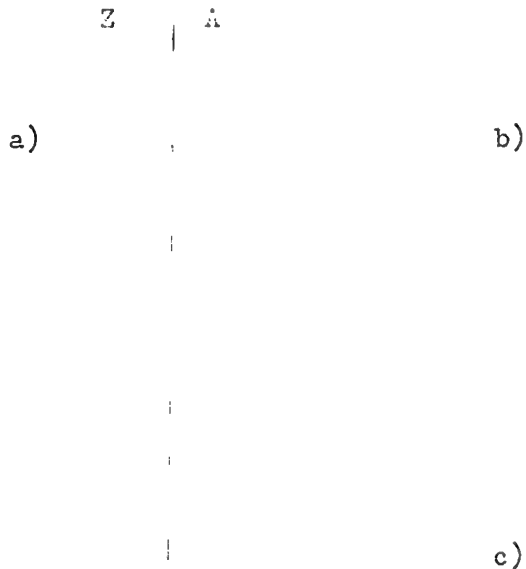


Fig.6-13 - A Directional Antenna Array Consisting of An

Exciter and a Parasitic Reflector $\frac{\lambda}{4}$ distant:

a) Relative Placement; b) Vector diagram; c) Directivity diagram along azimuth

exciter, is in phase with the current \vec{I}_a in the exciter. By traversing a distance equal to $\frac{\lambda}{4}$, this field, before driving the reflector, undergoes a 90° change in phase in the direction of lag (\vec{H}_{az}). On excitation by this field, current \vec{I}_z is induced in the reflector. By Lenz' Law, the direction of these currents is such that

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the field \bar{H}_{zz} they create, is, in the vicinity of the reflector, opposite in direction (antiphasal) to the field \bar{H}_{az} exciting the current in the reflector.

As the vectors \bar{H}_{zz} and \bar{I}_z coincide in phase, like the vectors \bar{H}_{aa} and \bar{I}_a , the currents in the reflector lag by 270° from the currents \bar{I}_a in the exciter or, what amounts to the same thing, are 90° ahead of the \bar{I}_a currents. The fields \bar{H}_{az} and \bar{H}_{zz} cancel each other out, being opposite in phase. Therefore, the array does not radiate behind the reflector, and the second major lobe of the horizontal pattern is cut out by the reflector.

The field \bar{H}_z lags by 90° in traversing the distance of $\frac{\lambda}{4}$ to the exciter, and, on reaching it, is in phase with the field \bar{H}_z , which it amplifies by 100%. Thus, the parasitic element reflects the energy radiated by the exciter, concentrating the total radiation of the system in the direction of the exciter alone. In so doing, the reflector makes the exciter more directional, narrowing the field intensity pattern by about 50% and reducing the radiated power density by $3/4$, approximately.

In practice, complete cancellation of the field behind the reflector does not occur because the currents induced in the reflector dipoles are 10 to 15% weaker than those in the exciter dipoles. Therefore, a portion of the energy seeps past the reflector and the pattern presents a small secondary horizontal lobe, although it is, however, incapable of affecting the work of the station. In practice, the reflector is so designed, by the use of induction bridges or other components, as to minimize the field behind the reflector.

The width of the primary horizontal beam of a multi-unit cophased array may be evaluated with the aid of the following equation:

$$\gamma^\circ \approx 51 \frac{\lambda}{b} \quad (6-7)$$

in which b is the width of the antenna,

λ is the operating wavelength.

Equation (6-7) is adequate for an antenna with a beam whose width does not exceed $25-30^\circ$.

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Often a flat metal grid, also $\frac{\lambda}{4}$ from the exciter, is used as reflector instead of a parasitic element (Fig.6-14). This type of antenna system is termed a curtain antenna.

The reflector also sharpens the vertical pattern of the system, eliminating radiation in vertical directions behind the reflector, and elongating the radiation in

Fig.6-14 - Antenna Array with Wire-Grid Reflector

the vertical directions in front of the exciter. As already noted in Section 6-1, the ground has a pronounced effect upon the shape of the vertical pattern. Ground reflection gives the vertical pattern a multi-lobular character, complicates the finding of targets at certain elevations, particularly low-flying targets at considerable distance from the location station.

Section 6-5. Director Antennas

Another type of directional antenna, widely used in stations operating in the very short wave and microwave ranges, is the director antenna.

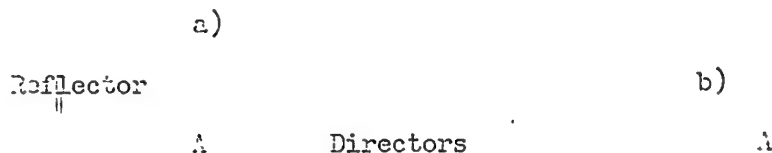
Director antennas are made up of wave ducts. Each wave duct consists of a system of dipoles, the centers of which are placed along the axial line of the duct (Fig.6-15). The transmitter drives a single half-wave exciter Λ of the usual folded type. The remaining dipoles are parasitic, their currents being set up by

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the electromagnetic field of the exciter.

The phase of the electromagnetic field of the exciter at a parasitic element is determined by the distance separating them. The e.m.f. induced in the parasitic element coincides in phase with that of the electrical field radiated from the exciter. The current set up in the parasitic element will differ in phase from that of the driving e.m.f. by an angle determined by the reactance factor in the input impedance



Exciter Support

Fig.6-15 - Wave Duct (a) and its Pattern (b)

of the parasitic element.

The magnitude and nature of the input impedance reactance of a dipole depends upon its length, measured in fractions of the operating wavelength. If the length of the parasitic element is a little over $\frac{\lambda}{2}$, the current in the element will lead the e.m.f. induced therein. Placed at a distance a little less than $\frac{\lambda}{4}$ from the exciter, the elongated parasitic element behaves like a half-wave parasitically-excited dipole $\frac{\lambda}{4}$ from the exciter, or, in other words, functions as a mirror. That is why the elongated parasitic dipole is called a reflector.

It may be shown by similar reasoning that the current induced in a somewhat shorter dipole will lag in phase behind the current in the exciter. As a result, the field induced by the shorter dipole is virtually in phase with the field radiated by the exciter, and amplifies its field. Thus, the shorter parasitic element guides the field of the exciter along the axis of the duct, and is therefore called a direc-

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tor. The superimposition of the dipole fields causes the wave duct pattern to have a width of 40 to 50° , becoming narrower as the number of dipoles in the duct increases. The axis of the pattern coincides with the axis of the duct (Fig. 6-15b).

A wave duct usually consists of one reflector and four or five directors in addition to the exciter. No exact method yet exists for analyzing a wave duct of any given number of dipoles. Many problems in this connection are resolved by experiment. The usual wave-duct design provides for a reflector length of 0.52 to 0.65λ , a director length of 0.3 to 0.46λ , a distance of 0.15 to 0.20λ from reflector to exciter, and a space of 0.2 to 0.25λ between adjacent directors.

Whatever is the length of the parasitic element, the point of induction of e.m.f. is located in its center. Therefore, wave duct dipoles are fastened firmly at their center points to an axial metal tube, and are not insulated therefrom, except for the exciter.

Single wave ducts are rarely used in radiolocation. Director antennas usually consist of 2 to 4-unit wave ducts, two to a tier. The exciters in a given tier must be driven in phase so as not to produce a pattern with two lobes in the horizontal plane.

An increase in the number of ducts within a tier narrows the horizontal pattern. The various tiers may be driven either in phase or counterwise. Switching tiers from cophased driving to antiphased raises the vertical pattern to an angle above the horizon. In circular scanning sets, this is employed to determine the elevation of the target, as a signal of identical strength, when feed is switched, is received only at a specific elevation. The height of the target may be determined by the elevation and the known range.

Section 6-6. Antennas with Parabolic Reflectors

The shorter the working wave, particularly as one enters the very short wave band, the more the methods of obtaining precision in direction approximate those used

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in optics. From optics we know that if a punctuate source of light be placed at the focus of a parabolic reflector, the reflector will rearrange the light rays into a parallel beam directed along the focal axis of the reflector.

In radiolocation systems, the "punctuate" source of radiation is usually a half-wave dipole placed at the focus of a parabolic reflector. The phases of all rays receiving parallel reflection attain identical values in any plane perpendicular to the focal axis of the reflector. Consequently, the shape of the radio wave radiated by

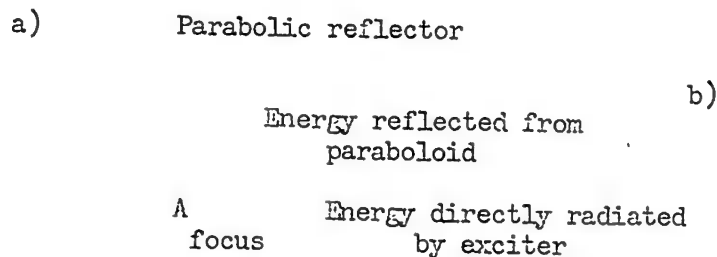


Fig.6-16 - Antenna with Parabolic Reflector:

a) Geometry of radiation; b) Antenna radiation pattern.

such an antenna is flat. Only those rays driven by the exciter directly along the focal axis, and not reflected, will be of different phase and will deform the wave front somewhat. To eliminate this undesirable phenomenon, the exciter is surrounded with a hemispherical screen to prevent direct radiation along the focal axis.

Figure 6-16 illustrates this type of antenna with a reflector in the form of a paraboloid of rotation. The body of the paraboloid may either be made of bent metal sheeting (solid or perforated) or of wire grid.

Despite the fact that all parallel rays are of identical phase in the plane of the front, their intensity will differ at various distances from the axis. This is explained by the fact that the individual points of reflection of the paraboloid are at different distances from the exciter. The intensity of the rays declines with

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distance from the focal axis. The concentration of the most intensive rays around the focal axis narrows the antenna radiation pattern and weakens the lateral lobes

a) b) c) d)

Fig.6-17 - Types of Parabolic Reflectors

- a) Parabolic cylinder; b) Full paraboloid; c) Truncated paraboloid;
d) Lemon-segment paraboloid

of radiation. The axis of the pattern coincides with the focal axis of the reflector.

When the reflector has the shape of a paraboloid of rotation, the width of the

Fig.6-18 - Typical Exciters

main beam (the flare of the pattern) may be calculated on the radiated power at half the maximum level (Fig.6-16b) by means of the following approximated formula:

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$$\gamma^{\circ} \approx 70 \frac{\lambda}{D} \quad (6-8)$$

in which D is the diameter of the largest cross-section of the paraboloid;

λ is the operating wavelength (in the same unit of measurement as D).

Equation (6-8) shows that the size of the reflector has to be 10 to 20 λ in order for the pattern to be sharpened. Use of the microwave band makes it possible to narrow the beam to tenths of a degree and to attain a power gain factor of more than 4,000 with paraboloids of appropriate size. An advantage of antennas with parabolic reflection is that the effect of the ground may be cut out almost completely due to the high directivity, assuming of course, that the sweep is not directly along the surface of the land or sea.

Sometimes truncated paraboloids, fruit-segment paraboloids or parabolic cylinders may be used. All these types of reflectors are illustrated in Fig. 6-17.

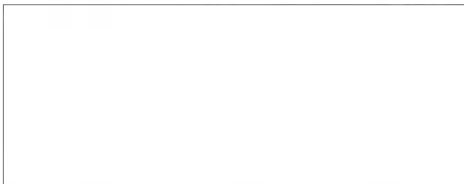
The parabolic cylinder has a cophased system of exciters along the focal line of the cylinder, and is used when the need is for a beam narrow along the flare of the reflector but wide along its length. When a parabolic cylinder is emplaced vertically, the result is a beam narrow in the horizontal plane (along the azimuth) and broad in the vertical plane (along the elevation). When a parabolic cylinder is in horizontal position, the result is a beam narrow along the elevation and wide along the azimuth.

In conclusion, let us note that in antennas of this type, the primary radiating element may be either an exciting dipole, a horn (Section 6-8) and systems thereof, or single-row slot antennas (Section 6-7).

Figure 6-18 illustrates typical exciters.

Section 6-7. Slot Antennas

Slot antennas are the invention of the Soviet scientist M.S. Neyman. A basic slot antenna consists of a slot in a driven cavity resonator through which electro-



magnetic energy from the resonator is leaked. The slot must be cut in such a manner that the electrical lines of force in the region of the resonator's electromagnetic

a) Surface of cavity resonator

Slot

Distribution
of intensity,
E, along length
of slot

c)

Distribution of
current, I,
along half-wave
dipole

b)

Fig.6-19 - Basic Slot Antenna

a) Notching of slot; b) Radiation pattern; c) Comparison with half-wave dipole

field are perpendicular to the length of the slot.

In order to attain effective directional radiation, the length of the slot must be half a wavelength in open space ($\frac{\lambda}{2}$), and the center of the slot must coincide with the point of maximum intensity of the electric field, in which case its distribution along the length of the slot (Fig.6-19a) will be analogous to the dis-

Fig.6-20 - Multi-Slot Single-Row Antenna

tribution of the current along the length of a half-wave dipole (Fig.6-19c).

For this reason, a slot of this type will behave like an equivalent half-wave dipole. However, unlike a half-wave dipole, the electrical field of which is paral-

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lel to its axis, that of a slot is transverse to the axis. Therefore, if a half-wave dipole has directivity only in the plane traversing its axis, a slot on the other hand is directional only in the plane perpendicular to that of the slot (Fig.6-19b).

Higher directivity is provided by a single-row multi-slot antenna, in which the fields of all the slots are kept in phase by making the distance between slots equal to a half-wave ($\frac{\lambda}{2}$) (Fig.6-20). Such an antenna behaves like a single-tier cophased array of half-wave dipoles, but yields its maximum radiation in a direction perpendicular to the plane of the slot. A single-row slot antenna is suited to utilization as the exciter of a parabolic cylinder reflector, being placed along its focal line with the slots facing inward.

Let it be noted that the type of slots we have discussed may also be cut into a waveguide.

Section 6-C. Horn Antennas

A waveguide through which electromagnetic energy is channelled, open at the end, may also be used as a directional source of radiation.

As already stated in Section 5-E, the ratio of the electric field intensity to the magnetic field intensity may be regarded as natural impedance not only in terms of the cross-section of a waveguide tube, but relative to the open space into which the energy flows from the end of the waveguide.

If the natural impedance of a waveguide at given frequency is determined by the type of channeled wave, and by its length λ in the waveguide (see eqs.5-15 and 5-16), then the open space offers the following resistance to the electromagnetic wave, which here necessarily acquires TEM structure:

$$Z_0 = \frac{\bar{E}}{H} = \sqrt{\frac{\mu_0}{\epsilon_0}}$$

In view of the fact that the magnetic permeability and dielectric constant of

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air are, respectively, of the following magnitudes

$$\mu_0 = 4\pi \times 10^{-7} \text{ [H/m]} \text{ and } \epsilon_0 = \frac{1}{36\pi} \times 10^{-9} \text{ [f/m]}$$

we obtain

$$Z_0 = 377 \text{ ohm.}$$

The most effective radiation from the end of a waveguide is that occurring while there is no reflection from the end of the waveguide. In order to bring the natural impedance of the waveguide into accord with that of open space, the cross-section of the terminal section of the waveguide is gradually expanded in such fashion as to permit this segment to play the role of a matching transformer. A portion of a waveguide designed for matching in this fashion is termed an electromagnetic horn.

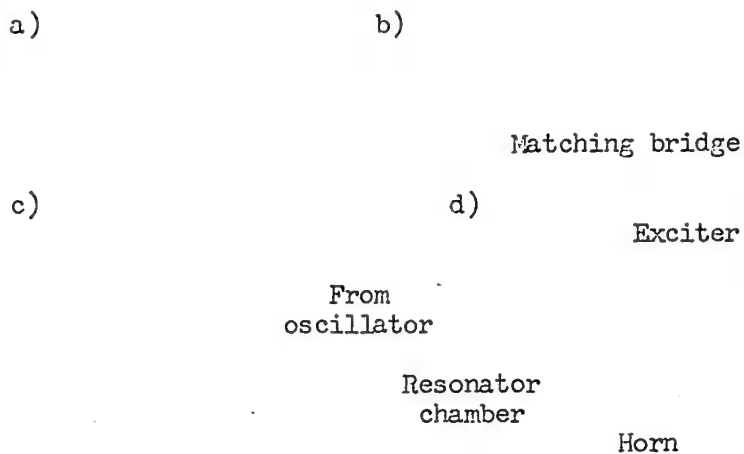


Fig.6-21 - Horn Antennas

a) Sectoral horn; b) Pyramidal horn; c) Conical horn; d) Driving by means of resonator chamber

The basic types of horns are illustrated in Fig.6-21. A horn provides maximum radiation in the direction of its geometrical axis. The radiation pattern of a horn is rendered more acute by widening the flare and increasing the length, as this

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creates optimum relationships between flare and horn length. A properly-designed horn may be considered to provide cophasality of the electromagnetic field at all points in the horn aperture, resulting in radiation of a wave with a flat front.

A horn antenna is of rather high directivity. However, its pattern includes not only the main lobe but side lobes, the presence of which is due to the fact that the electromagnetic oscillations at various points of the aperture are not perfectly in phase. Horn antennas find their main use in radiolocation as exciters of parabolic reflectors. The exciter may contain either one or a system of horns along the focal line of the parabolic cylinder.

Let it be noted that not only waveguides may end in horns, but cavity resonators driven in the usual way by the excitation of waveguides, (Fig.6-21d).

Section 6-9. Dielectric Antennas

Electromagnetic energy may be channeled not only in waveguides of air in metal walls, but in dielectric rods surrounded by air. In the latter instance, the outer surface of the rod will also serve as the boundary layer between two media with different electrical characteristics. Therefore, the mechanism for the propagation of radio waves along a dielectric rod will be identical in principle with the mechanism of propagation in a metal-walled waveguide. The only difference is that complete internal reflection from the boundary surface between the media is capable of occurring in the given instance only at supercritical frequencies. At subcritical frequencies, the radio-wave energy is reflected from the contact surface only in part, and is therefore distributed along the dielectric waveguide, while in part it penetrates through the boundary layer into the surrounding space, being refracted in the process. Dielectric antennas take advantage of the refraction of radio waves on passage into the air from the dielectric.

A dielectric antenna (Fig.6-22) consists of a cone with a rounded end, made of a weakly absorbing dielectric (usually polystyrene). The conical shape of the radi-

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ating dielectric provides the smoothest possible transition from the greater natural impedance of the rod to the lesser natural impedance of open space. This weakens reflection from the rod cone and provides radiation of good effect.

The directivity of the dielectric antenna is governed by the phase difference

- a) Metal chamber b) Cone of weakly absorptive dielectric

Exciter Cross section of cone

Fig.6-22 - Dielectric Antenna

a) Design; b) Radiation pattern

of the electromagnetic fields radiated by different points at the surface of the dielectric rod, and by the difference in path of these fields to the target point in space. The maximum radiation is in the direction of the geometrical axis of the rod.

A rod of considerable length and large diameter is needed to render the pattern more acute. Let us note that as the length of the rod is increased there is a simultaneous improvement in the sharpness of the major lobe but also an increase in the prominence of the side lobes.

Added directivity is attained by means of polyrod antennas set up side by side at intervals equal to a whole number of waves so as to assure driving in phase, with the use of a branched system of coaxial lines.

Section 6-10. Lens Antennas

The principle of operation of the electromagnetic lens is based on the fact that the phase velocity of radio wave propagation in a dielectric is greater by a

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factor of $\sqrt{\epsilon}$ than its velocity in a cavity. In antenna engineering, lenses are used to attain cophasality of the field at all points in the antenna output aperture, thus simplifying the problem of assuring a proper radiation pattern with an exciter not very deeply positioned.

Thus, if a short horn is used in a horn antenna, it is difficult to attain a narrow pattern even when the flare is wide. The difficulty arises from the fact that under these conditions the wave front at the horn outlet will be spherical and not flat, and the phase of the oscillation will differ from point to point at the outlet aperture, due to the fact that at the edges it is farther from the exciting waveguide than at the middle. The result of the variation in phase of the oscillations

Horn

Dielectric lens

a)

b)

Fig.6-23 - Lens Antennas

a) Dielectric lens with components of equal width and rear profile forming special pattern; b) Metal lens of waveguide elements of equal length, but differing in width as designed

radiated by various points at the outlet aperture is to cause the radiation pattern to reveal a deep valley along the geometrical axis of the horn.

To eliminate the axial trough a built-up dielectric lens is inserted in the outlet aperture of the horn (Fig.6-23a). The rear profile of this assembly is so

designed that the outside rays travelling the longest path in the horn flare are accelerated in the dielectric exactly to the degree necessary to reach the outlet aperture in phase with the middle rays travelling the shortest path.

In view of the high cost of the dielectric lens, a metal lens is often used instead. In the latter case, the assembly of dielectric elements is replaced by one of waveguide components (Fig.6-23b). As we have seen in Chapter V, the phase velocity of a wave in a waveguide is always greater than the phase velocity of air. Therefore, an assembly of waveguide components with rear profile of special design is equivalent to a dielectric lens.

The H_{10} is the type of wave most generally employed. The phase velocity of such a wave in a waveguide is the greater, the less the distance between the walls. Therefore, the length of all the waveguide components of the lens may be regarded as identical, but the components used at the extremes of the horn flare are narrower than those at its center (Fig.6-23b). In this case, the extreme rays in the horn flare will be accelerated more than those in the center. The oscillations at all points in the outlet aperture will be in phase, and the purpose of the electromagnetic lens will thus be fulfilled.

Use of a lens permits reduction by a factor of 20 to 25 in the length of the horn required to attain the necessary acuteness of the antenna radiation pattern.

Section 6-11. Operating a Transmitter and a Receiver on a Single Antenna

As already stated in Section 3-2, the pulse method permits a single antenna to be used in a radar for sending and receiving. The reflected pulses are received in the intervals between the transmitted pulses, when the transmitter is not working.

The need for only one antenna considerably reduces the weight and size of a radar and simplifies its high-frequency channel.

Switching the antenna from receive to transmit is done automatically by means of an antenna transmit-receive switch, performing two main functions:

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a) connecting the antenna to the transmitter output during the period in which the uhf pulse is generated, with simultaneous shorting of the receiver input, and switching it off the antenna with the purpose of protection from the effects of the powerful outgoing pulse;

b) unlocking the receiver input and connecting it to the antenna for the entire interval between the outgoing pulses, while cutting out the sender output.

In order for these functions to be performed properly, the antenna transmit-receive switch should fill the following requirements:

1. The duration of "transmit-receive" switching should be tenths or hundredths of the duration of the operating pulse, meaning that switching should be without lag and virtually instantaneous. Any significant delay in switching the receiver off the antenna relative to the moment at which oscillation of the outgoing pulse begins, has the effect, to begin with, of endangering the receiver input, thus requiring supplementary protection, while, secondly, it has a negative effect upon the channeling of the energy from the transmitter into the antenna in view of the fact that some of it is diverted into the receiver channel.

On the other hand, any considerable delay in connecting the receiver in terms of the moment at which pulse oscillation ends, enlarges the small "unscanned" area, while lagging switching of the transmitter off the antenna worsens the channelling of energy from the antenna into the receiver due to diversion into the output circuits of the transmitter.

2. Switching the antenna from receive to transmit and vice versa should not result in a mismatching of uhf channel components.

3. The commutator details performing the switching of the antenna should themselves consume minimum energy, so as not to induce any significant reduction in the efficiency of the uhf channel of the station.

In stations operating in the very short wave range, switching of the antenna from receive to transmit is performed as a rule by means of quarter-wave and half-
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wave feeder segments, various types of gas-discharge arresters being used as the commutator devices. If a voltage equal to or exceeding the ignition voltage be applied to the electrodes of the gas-discharge arrester, the ionization processes developing almost instantaneously cause the arrester to break down. The resistance of the broken-down arrester is very small and may be regarded for all practical purposes as a shorting of the points between which the arrester is cut in. On the other hand, if the voltage applied to the arrester should drop sharply by compari-

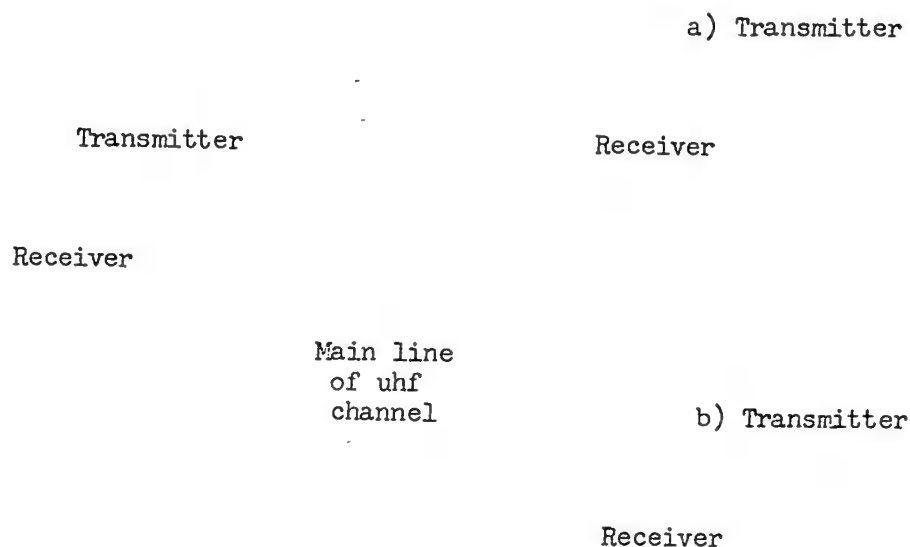


Fig.6-24 - Equivalent Circuit of Transmit-Receive Switch

a) For transmit operation; b) For receive operation

son to its ignition voltage, the arrester will be extinguished. The resistance of the arrester when not broken down is very high and may be taken to be infinite.

The receive-to-transmit switching circuits used in the antennas of various stations differ, but they may often be reduced to the equivalent circuit shown in Fig.6-24. At the instant when generation of the transmitter pulse in the branched system of the uhf channel begins, a traveling voltage wave develops which simultan-

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ously breaks down both arresters. The input resistance of the quarter-wave line connected at points dd is equal to infinity when arrester GP-1 is broken down, and this is the same as switching this segment out of the uhf channel (Fig.6-24a).

Therefore, during the period when the pulse is being generated the transmitter is connected to the main line of the channel to the antenna. At the same time, the input resistance of the half-wave line connected to points nn is zero when the arrester GP-2 is broken down, which is equivalent to shorting the receiver input. If the receiver input at points nn is shorted, the input resistance at points aa on the quarter-wave line to the receiver is equal to infinity, which is equivalent to disconnecting that line (and consequently, the receiver with the shorted input) from points aa on the main uhf channel, for the entire generation period of the outgoing pulse.

Thus, during the generation of the outgoing pulse, only the transmitter is connected to the antenna.

At the instant when generation is concluded, the wave of voltage traveling from the transmitter damps down and both arresters are temporarily extinguished. Now the input resistance of the quarter-wave line, already open at the end, will be equal to zero, which means short-circuiting at points dd (Fig.6-24b) and disconnecting the transmitter line from points aa on the main line. At the same time, the line impedance of the half-wave line at points nn on a half-wave line from an arrester that has not been broken down, GP-2, will be infinite, which is the same as saying that there is no half-wave line, that the receiver input has been switched off and connected to points aa on the main.

Therefore, during the interval between outgoing pulses, when the arresters are not broken down, it is only the receiver, ready to receive reflected signals, that is connected to the antenna.

A fundamental inadequacy of the circuit in Fig.6-24 is that the arrester, GP-2, acts directly on the receiver input, which is therefore directly subject to the ef-
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fect of the traveling-wave voltage peak breaking down the arrester. After breakdown, however, as a result of which the voltage drop at the arrester electrodes GP-2 becomes small, the receiver input will be protected against the effects of the transmitter impulse. An additional arrester is usually provided at the input circuit of the receiver for the protection against the breakdown peak.

At shorter wavelengths, starting with the decimeter range, in which the first component of the receiver input is usually a crystal mixer very sensitive to overloads, protection of the receiver becomes particularly important. In this connection

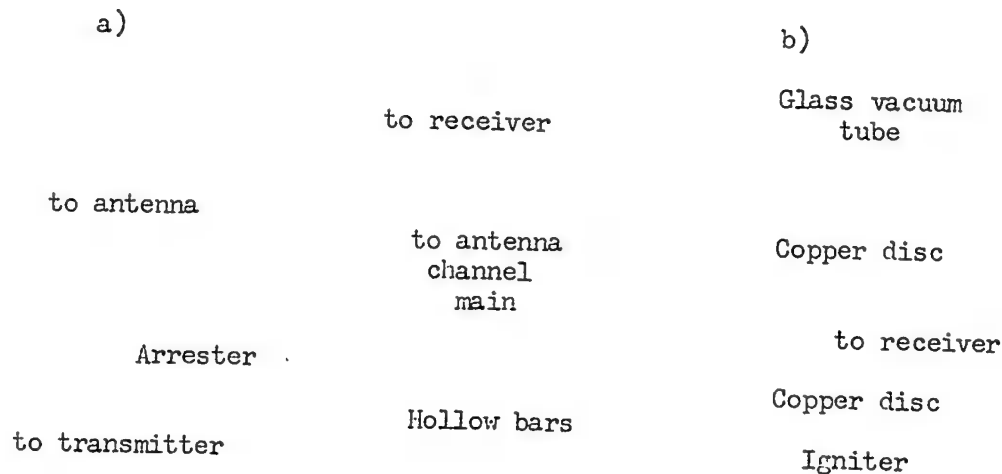


Fig.6-25 - Tuned Transformers in Receiver Protection Circuits

a) Coaxial-cylindrical cavity; b) Cavity resonator with vacuum arrester

it is important not only that there be low voltage on the disrupted arrester, but that puncture be easier, i.e., that it break down before the voltage traveling along the line from transmitter to antenna attain its full pulse magnitude. With this object in mind, the line to the receiver input is connected via a cavity resonator acting as a tuned voltage transformer.

In the microwave band a coaxial-cylindrical cavity is used to perform the function of tuned transformer, (Fig.6-25a). It consists of a coaxial half-wave line shorted at both ends. A standing voltage wave builds up in this type of resonator,

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the minima being at the shorted ends, and the maxima in the middle where the arrester is emplaced. Thus, the arrester is in a loop where the voltage is always higher than in the line from the main to the resonator. The resonator behaves like a voltage step-up transformer. The arrester is disrupted before the traveling-wave voltage in the line attains breakdown magnitude. Puncture of the arrester changes the parameters of the resonator, taking it out of resonance, with the result that the input resistance of the resonator for the receiver line connected close to the shorted end, falls sharply to a very low level. In addition, the receiver line remains in the voltage node. This provides good receiver input protection.

On reception of signals reflected from the target after generation of the transmitter pulse has ceased, the arrester is not disrupted, and the coaxial-cylindrical cavity now behaves like a half-wave resistance transformer matching the receiver line characteristic to that of the line from resonator to main. This assures a high traveling-wave characteristic when the energy of the reflected signal is channeled from antenna to receiver.

At the beginning of the very short wave band, when use of coaxial lines is still possible, a hollow-space cavity resonator is used as tuned transformer (Fig.6-25b). This device has a higher Q-factor and consequently higher resonant properties than a coaxial-cylindrical cavity. Magnetic loops are used for connection with coaxial branch lines.

The arrester is designed as a separate, replaceable component enclosed in a gas-filled glass tube. This type of arrester is often termed a gas-discharge tube or soft rhumbatron. Some gas-discharge tubes have an igniter as well, connected to an auxiliary constant-voltage low-power source of 700 to 1000 volts. The purpose of the igniter is to provide the first-stage ionization necessary to reduce the arrester breakdown time to approximately 0.01 microsecond, and to provide rapid recovery in the interval between outgoing pulses, in readiness for disruption by the next transmitter pulse.

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The switch cavity resonator can be connected with the receiver by a coaxial line and, on the main-line end, with the waveguide by a slot.

Many waveguide transmitter-receiver antenna switch systems are available. Most of these use twin-T waveguide connections, consisting of four waveguides in a single junction (Fig.6-26). Three lie in the plane of the wide edge, and the fourth is perpendicular thereto. The TE_{10} wave from the waveguide A can only reach waveguides B and C, since the direction of the electrical lines of force in all these waveguides is identical: parallel to the narrow edge. This wave cannot enter the waveguide D_1

a)

b)

c)

Fig.6-26 - Twin-T Connection

a) General view; b) Energy moving simultaneously from waveguides B and C enters waveguide D; c) Energy fails to enter waveguide D, passing it by

since the electrical lines of force penetrating into the cross-section of the waveguide D pass along it, which does not conform with the structure of TE_{10} . However, the TE_{10} wave cannot turn from A to D at a right angle, since the electrical lines of force in D would be parallel to its wide edge, which in waveguide D is equivalent to a conversion of TE_{10} to TE_{01} wave. The critical length of the wave TE_{01} , unlike that of TE_{10} , is determined by the narrow side rather than by the broad side of the waveguide D.

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The dimensions of the double T-section connection are designed so that the oscillator wave driving the waveguide system will be supercritical for the TE_{01} mode. This makes it impossible for the TE_{01} wave to be propagated in the waveguide D. Thus, no diversion of energy from waveguide A to waveguide D can occur.

For the same reason, a wave coming in along waveguide D cannot penetrate waveguide A. However, it can enter waveguides B and C, as the turning of the wave in the plane of the electrical field unaccompanied by changes of the corresponding cross-sectional dimensions of the waveguide, is not accompanied by transformation of the TE_{10} wave into a TE_{01} wave.

A wave entering the node along waveguide B may enter waveguides C or A, as the electric lines of force in all these waveguides run along the narrow side.

A wave intersecting the branch in waveguide D is incapable of entering it on spread from waveguide B to waveguide C or vice versa only if on both sides of this branch the electric lines in waveguides B and C are in opposing directions (the electric fields differing 180° in phase). This is necessary if the electric lines

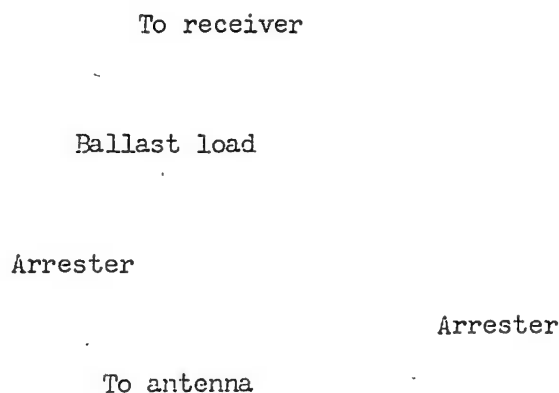


Fig.6-27 - Employment of Twin T-Section Connection in Waveguide Channel of Station in Microwave Band

are to turn 90° , and is depicted in Fig.6-26b.

Similarly, waves moving toward the node simultaneously along waveguides B and STAT

may branch into waveguides D or A only if they are in opposite phase on reaching the opposite side of that branch. However, if they are identical in phase on attaining opposite sides of the branch, their fields, turned by 90° , will, at the cross-section of the branch, be directed toward opposing sides and will cancel each other out (Fig.6-26b). In this case, no branching of the energy is possible.

Let us examine the circuit in Fig.6-27 as an example of the utilization of a double T-section connection in a waveguide system for a transmit-receive switch. When the transmitter is in operation, the waveguide wave 4 which it radiates branches along waveguides 2 and 3 and breaks down the arresters. Waves 6 and 5, reflected by the arresters, traverse the full routes $2\frac{\lambda}{4}$ and $2\frac{\lambda}{2}$, which differ by $\frac{\lambda}{2}$, on their return to waveguide 1, and consequently reach waveguide 1 in antiphase. Therefore, the energy from the transmitter (waveguide 4) freely penetrates the antenna (waveguide 1). The insignificant leakage of waves past the arresters is directed through waveguides 7 and 8, following identical paths from waveguide 4 to waveguides 9 and 10, and are in phase on arrival there. These leakage waves are not able to penetrate into the receiver (waveguide 10). They branch only into waveguide 9, the end of which is terminated at matched ballast impedance, where complete absorption of the energy of the leakage waves occurs. When the arresters are not disrupted, the wave of signal received from waveguide 1 is unable to penetrate waveguide 4 (the transmitter), but branches into waveguides 2 and 3. The lines of force of the electric field make a 90° turn in opposite directions on entry into these waveguides from waveguide 1. Consequently, on leaving waveguide 1, the electric fields in waveguides 2 and 3 differ in phase by 180° . Continuing along waveguides 8 and 7, they enter waveguide 10 (the receiver) in opposing phase, and are therefore able to penetrate it freely.

The circuit just described provides highly satisfactory protection for the receiver, and may be used for very powerful stations in the microwave band, in view of the employment of waveguides.

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Section 6-12. Operation of Transmitter and Receiver on Separate Antennas

In certain types of radiolocation stations, the transmitter and the receiver are connected to their own antenna array. In this design there is usually no need for a "transmit-receive" switch.

To weaken the reciprocal distortions of radiation patterns, the receiving and transmitting components of a station and their antennas sometimes have to be placed tens of meters apart. This increases the space requirement for a station.

Another major weakness of the use of separate antennas for transmitter and receiver is the bulkiness of the stations and the extreme weight involved, particularly as a single transmitting and a single receiving antenna are not always adequate. In distant warning stations, for example, effective following of targets (particularly low-flying targets) at long range requires a narrower radiation pattern than is the case with short-range stations. This requires either the maintenance of two antennas (one for long-range, and one for short) or switchover to the necessary radiation pattern on switchover from one range to the other.

On the other hand, in order to combine rapid detection and accurate bearings, certain types of stations conduct their searches in two stages: coarse-range search with a wide-beam antenna, and the taking of bearings and target following with another, narrow-beam antenna. In this arrangement, simultaneous azimuthal and tilt direction finding are sometimes conducted in yet another way. One antenna, the beam of which is narrow enough in the vertical plane and wide enough in the horizontal to facilitate the azimuth search, is used for elevation determination. Azimuthal direction finding is conducted by means of another antenna whose beam is, on the other hand, narrow enough in the horizontal plane and, to facilitate determination of tilt, wide enough in the vertical.

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CHAPTER VII

METHODS OF TAKING BEARINGS ON TARGETS AND OF SWEEP SCANNING

Section 7-1. Methods of Taking Bearings on Targets

Methods of taking bearings are used in radiolocation to determine the angular coordinates of targets. The measured elevation or azimuth is calculated on a specific antenna position such as to guarantee that the reflected signal will be of normal magnitude corresponding to the direction to the target.

Three principle methods of direction finding are used in radiolocation:

- a) Bearings by maximum signal strength;
- b) Bearings by the zone of equal signal strength;
- c) The determination of bearing by phase.

In the majority of cases, a single transmit-receive antenna is used, automatically switching to transmission and to reception at the pulse repetition rate.

Search by electromagnetic beam and target bearing determination is performed by moving the antenna. The close relationship between the method of direction finding used at a given station and the method whereby the electromagnetic beam travels through space (scanning) and the type of indicator used is particularly clear when a single antenna is used.

Section 7-2. Taking Bearings by Maximum Signal Strength

The radiation pattern, $e = f(\alpha)$ of the receiver antenna describes the relation-

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ship of the voltage e at the input of the receiver attached to the given antenna, to the angle between the base line for angle measurement and the direction of the reflected signal. The length of the vector radius of the pattern in the direction of the incoming signal determines the signal voltage at the receiver input for the

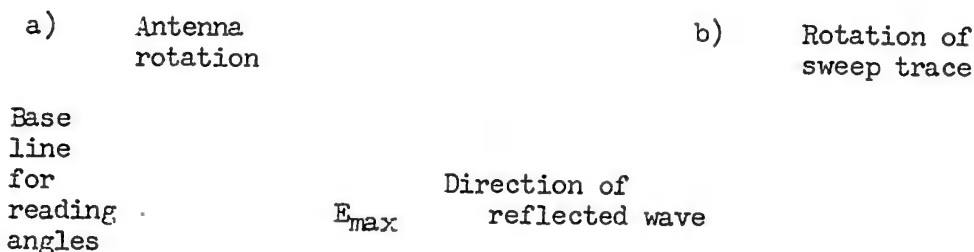


Fig.7-1 - Taking Bearing by Maximum Signal Strength

a) Position of rotating pattern at moment bearing is taken; b) Indicator image at the moment when bearing is taken.

given direction. When the angular position of the antenna is changed, its pattern revolves around the center point (Fig.7-1a). In direction finding by maximum signal strength, the bearing angle (α_n) is read at the moment when the direction of the incoming reflected signal coincides with that of the maximum vector radius E_{max} of the pattern, as this means that the signal reflected to the receiver input will be at its maximum. The function of the angular coordinates indicator in this situation is to fix the moment when the reflected signal is at its maximum and to permit the reading of the corresponding maximum angle of the antenna. As already indicated in Section 3-2, the sweep trace on the indicator screen is caused to rotate about its center (Fig.7-1b), so as to imitate precisely the angular positions of the antenna. In this case, the reflected signal determines the brightness of the pip. In reading

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bearings by maximum signal strength, the pip is at its brightest at the moment when the bearing is taken.

The accuracy with which angular coordinates are measured depends upon the sensitivity of the indicator to change in reflected signal strength and on the sharpness of the pattern of the receiving antenna for taking bearings. Figure 7-1a shows that at angles approaching that of maximum strength in the pattern, changes in the vector radius are quite small. In the vicinity of the maximum, the sharpness of the pattern is at a minimum. Reading bearings by maximum signal strength is not a highly accurate method, but it does facilitate the discovery of weak signals against receiver background noise, and is used in distant-warning and circular-scanning stations in which the problem is to read the bearing on a distant target, but where high accuracy is not required.

Section 7-3. Determining Bearings by Zone of Equal Signal Strength

Let us assume that we have two identical antennas (for example, two wave

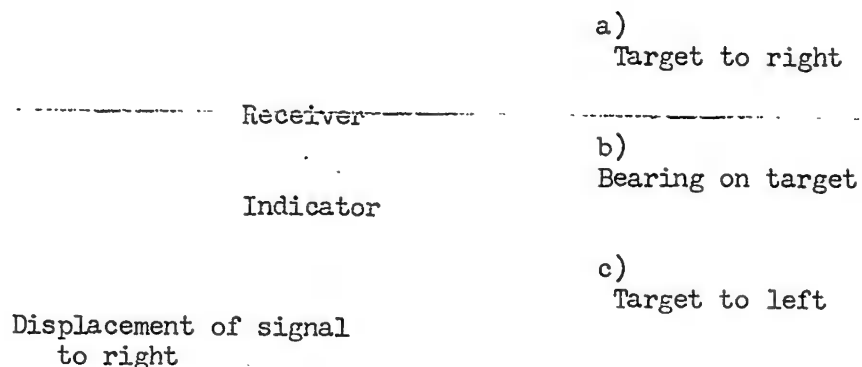


Fig.7-2 - Taking Bearing by Zone of Equal Signal Intensity

ducts) so arranged that the axes of their patterns intersect at a given angle. This STAT

half a degree, which is far superior to that attainable by the method of maximum signal strength.

Taking bearings by the equal-signal-zone method makes use of the voltage differential between e_1 and e_2 , which changes far more sharply than these two voltages themselves. This permits the utilization of antennas with wide patterns and, consequently, of small size. However, the range is somewhat less than that attainable in range-finding by maximum signal strength, as the signal is below its maximum in the direction of the equal-signal-strength zone.

Section 7-4. Taking Bearings by Phase

Another method permitting the taking of high-accuracy bearings on the azimuth, for example, is the phase method. It has found application in stations in the very short wave band.

In the phase method, the antenna array consists of three co-linear horizontal half-wave dipoles (Fig.7-3), d being the distance between centers.

LB CB PB

 FY

 AP

to receiver

Fig.7-3 - Schematic of Phase Method of Taking Bearings

LB) Left dipole; CB) Middle dipole;
PB) Right dipole; FY) Phasing device;
AP) Antenna switch

the middle dipole, by an angle of $\varphi = \frac{2\pi}{\lambda} \Delta r$, and the e.m.f. \bar{E}_n induced in the

Let the wave reflected by the target impinge from the right, causing its rays to form an angle θ to a line perpendicular to the axial line of the dipoles, resulting in a difference in the paths of the rays to the extreme and middle dipoles $\Delta r = d \sin \theta$. The e.m.f. \bar{E} , induced in the left dipole, will lag in phase behind that, \bar{E}_c , in

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right dipole will lead \bar{E}_c by the same angle φ (see the vector diagram in Fig.7-4a).

The e.m.f.s of the outer dipoles are delivered along lines of equal length to the output of the phasing device FY, and the e.m.f. of the middle dipole to the input of the antenna switch AP. The phase differences between the e.m.f. at the phasing device input and at the antenna switch input remain the same as they were at the dipoles themselves.

The phasing device FY rotates the phase of the e.m.f. of one of the outside dipoles (say, the right dipole) by 180° , and then superimposes it upon that of the other outside dipole (the left). It constitutes a segment of a fixed two-wire line $\frac{\lambda}{2}$ long (Fig.7-5). Sliding contacts permit the energy to enter the antenna-switch feeder. The sliding contacts are placed on a bridge in such a position (aa) that the path of the wave traveling from one of the outside dipoles to the sliding contacts is longer by $\frac{\lambda}{2}$ than that of the wave from the other outside contact, which is equivalent to obtaining a 180° phase difference between these waves at the out-

a)

b)

c)

Not on target, antenna
must be turned to left

On target

Not on target, antenna
must be turned to right

Fig.7-4 - Vector Diagrams of Phase Method

put of the phasing device (points aa). Thanks to this phase shift, the resultant

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e.m.f. E_F at the output of the phasing device, now lags by 90° behind the e.m.f. \bar{E}_C of the middle dipole (Fig.7-4a). As a result of the passage through the feeder (see Fig.7-3), this e.m.f. now lags in phase by another 90° , and enters the antenna

Phasing bridge

To antenna switch

Fig.7-5 - Phasing Device (Bridge)

switch input AP (already designated as \bar{E}_F , for ease of reference) in antiphase to the e.m.f. \bar{E}_C .

In position I, the antenna switch superposes \bar{E}_F and \bar{E}_C , while in position II it subtracts them. The resultant signal emitted from the antenna switch output attains the receiver input as follows: as $\bar{E}_C - \bar{E}_F$ at position I, and as $\bar{E}_C + \bar{E}_F$ at position II. Therefore, if the signal comes in from the right, the indicator screen will show it to be smaller in position I (due to the fact that e.m.f.s \bar{E}_C and \bar{E}_F are in opposite phase) than in position II. If the wave reflected by the target has arrived from a direction to the left of the perpendicular to the axial line of the dipoles, the e.m.f. \bar{E}_F would change its polarity, and the relationship between the amplitudes of the signals on the indicator screens would be the opposite (Fig.7-4c).

In order to take a bearing, the antenna array is rotated in the horizontal plane until the signals on the indicator screen are equal in both positions of the antenna switch. This requires that the direction of the reflected wave be perpendicular to the axial line of the dipoles. When this occurs there will be no difference in the paths of the rays, the e.m.f.s of all three dipoles will coincide in phase, and the resultant e.m.f. E_F of the phasing device will be zero (Fig.7-4b). In STAT

will cause the patterns to overlap (Fig.7-2). An antenna switch AP in constant operation automatically alternates the receiver between the two antennas. When the antennas and the direction of the reflected target beam are each in a given position, the receiver input voltages e_1 and e_2 will differ in accordance with whether the antenna switch is at position I or II. Joint rotation of the antennas causes rotation of both pattern outlines (the solid line and broken-line), with a corresponding change in the e_1 and e_2 voltages. There will be a change, accordingly, in the relationship between the amplitudes E_1 and E_2 of the signals on the indicator screen. When the antennas are in a position in which the direction of the reflected beam is that shown by line OA, the voltage on the receiver input will be equal when the switch is in either position ($e_1 = e_2$). At that instant, the amplitudes of the signals E_1 and E_2 on the indicator screen will also be equal. The direction to OA is termed the equal-strength zone, and the angle between that zone and the direction of the reflected ray gives us the bearing of the target.

Thus, in taking bearings by the zone of equal signal strength, the angle being measured is read at the moment when the signals E_1 and E_2 on the indicator screen are equal in amplitude at both positions of the antenna switch (Fig.7-2b). In order to facilitate comparison of the amplitudes, the signal received when the antenna switch is in each of the two positions is automatically shifted to and reproduced independently on a particular side of the screen (on the left for position I and the right for position II), but both are seen simultaneously because of the speed of switching and afterglow on the indicator tube. It should be noted that in the given case the indicator tube is not a reading device but merely an indicator for the instant when the bearing is correct, providing information as to when a reading of angular coordinates should be taken. This reading is taken on a dial, the pointer of which is connected to the antennas by automatic devices, and controls their position.

In practice, the equal-signal-zone method readily provides accuracy to within

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both positions of the antenna switch, only the e.m.f. \bar{E}_c will reach the receiver input. When the antenna is in this position, an azimuth bearing has been obtained. The azimuth of the target may be read off at this moment on the dial of the indicator showing the position of the antenna. The indicator tube serves only to reveal the

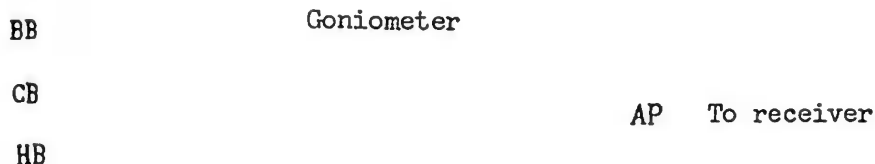


Fig.7-6 - Schematic of Goniometric Method of Determining the Angle of Elevation

BB) Upper dipole; HB) Lower dipole; CB) Middle dipole; AP) Antenna switch

moment when the bearing has been obtained, as in taking bearings by equal-signal-strength zone.

Thus, the phase method of taking bearings comes down to equalization of the phases of the e.m.f.s induced in the antenna dipoles. This is why it is called the phase method.

A second example of the employment of the phase method of direction finding in stations in the meter band is measurement of elevation by an ultrashort wave

goniometer (Fig.7-6). The upper and lower dipoles of the antenna system employed for this purpose are at different altitudes above the earth (say at $2\frac{\lambda}{2}$ and $3\frac{\lambda}{2}$). The vertical patterns of these dipoles will differ. As already indicated in Section 6-1, the number of lobes in the vertical pattern is equal to the number of half-waves that can fit into the distance between the dipole and the ground. In our example, the vertical pattern of the upper dipole BB has three lobes, and that of the lower dipole HB has two.

The goniometer consists of a system of three coils, two fixed (stators) and one moving (rotor), capable of rotating within the stator coils. The upper and lower

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dipoles are connected to the two mutually perpendicular stator coils by means of two-wire feeders. The rotor coil of the goniometer is connected to the antenna switch by means of the feeder.

Due to the difference in elevation above the ground at any tilt of the target, the e.m.f.s induced in the upper and lower dipoles, and the currents in the stator coils of the goniometer, connected therewith, will differ in amplitude and phase. The fields of the stator coils induce voltages at the ends of the rotor coil which differ in amplitude and phase, in the general case.

However, at any elevation, it is possible to find a specific position of the goniometer rotor coil for which the voltages induced at its ends by the fields of the two stator coils will be equal in amplitude and opposite in phase. In this case, the resultant voltage at the ends of the rotor coil will be zero.

The input of the antenna switch AP receives not only the resultant voltage of the goniometer rotor coil, but the e.m.f. induced in the middle dipole CB. When the antenna switch is thrown from one position to the other, the e.m.f. of the middle dipole is either added to or subtracted from the goniometer voltage. If the position of the rotor coil has not been properly adjusted, the signals on the elevation indicator screen will differ in value as between antenna switch positions I and II (as in Fig. 7-4a and c).

The rotor coil is turned until the signals are equal when the switch is in both positions, indicating a bearing on the target. The elevation at this moment is read off on the goniometer dial (scale).

Section 7-5. Methods of Sweeping the Electromagnetic Beam

Sweeping, or scanning, is the name given to moving the directional electromagnetic beam of the transmitting antenna for consecutive radiation of surrounding space.

There are quite a large number of methods of scanning, but all are variant. STAT

the four main types. These four main types are:

- a) circular scanning;
- b) screw scanning;
- c) conical scanning;
- d) spiral scanning.

As already noted in Section 7-1, the method of scanning used, is coordinated with the method of taking bearings employed in the given station, and the type of indicator of angular coordinates. On the other hand, the rate at which the electromagnetic beam is scanned, its width, and the pulse repetition rate, must always be coordinated.

Section 7-6. Circular Scanning

In circular scanning, the electromagnetic beam and the antenna are rotated in

Horizontal plane

Antenna

Fig.7-7 - Circular Scanning

a horizontal plane around a vertical axis with an angular velocity of $\omega(\varphi)$ radian per second (Fig.7-7).

If the width of the horizontal pattern is $b(\varphi)$ radian, the time required for a circular sweep of the beam across the entire width is $\frac{b(\varphi)}{\omega(\varphi)}$ second. A located target is within the range of the beam during that period. But in order to obtain a satisfactory target pip on the indicator screen, a minimum number of pulses n determined by the type of indicator, must be directed at the target during this period.

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Therefore, when circular scanning is used, it is necessary to consider, in determining pulse frequency, that it must exceed the minimum required by that type of scanning:

$$F_{u \text{ min.}} = n \frac{\omega(\varphi)}{b(\varphi)} \quad [\text{pulses/sec}] \quad (7-1)$$

Circular scanning is used in distant warning stations and in 360° search stations employing direction finding by maximum signal to measure the target azimuth. The indicator image obtained by circular scanning is that which may be seen in Fig.7-1. The rate of rotation in circular scanning is usually small, constituting 4-6 rpm., and the pulse rate does not exceed 500 per second.

A special type of circular scanning is sector scanning, when rotation of the

a) b) Target pip

Fig.7-8 - Sector Scanning for Elevation

antenna is replaced by motion to and fro within a limited sector controlled by the electromagnetic beam.

Combination of sector scanning and the taking of bearings on maximum signal strength may be employed to measure the elevation of the target by means of a special antenna operated by a transmit-receive switch. The antenna and its beam are rocked to and fro within a specific interval of tilt, say from 0 to 45° (Fig.7-8a). The rotatory motion of the scanning trace on the elevation indicator screen may, in the given case, conveniently be replaced by parallel up-and-down motion (Fig.7-8b). Each angular position of the antenna corresponds to a specific position of the scan-STAT

ning trace on the elevation indicator screen. The brightness of the pip on the screen is at its maximum when the elevation is on target.

Equation (7-1) remains valid when sector scanning is used. When the elevation is to be measured, it is necessary to write into the equation the beam width b_θ for the elevation, and the angular velocity ω_θ of antenna motion in the vertical plane.

Section 7-7. Screw Scanning

In screw scanning, the electromagnetic beam emitted by the antenna is periodically rotated through a given azimuthal sector, with simultaneous but considerably slower change in tilt. By its simultaneous participation in these two types of motion, the beam describes a distorted screw-like line in space (Fig.7-9a).

Screw scanning is also usually combined with direction finding by maximum impulse. In screw scanning, the indicator permits simultaneous measurement of two angular coordinages, the azimuth φ , and the elevation θ (Fig.7-9b). The vertical motion of the scanning trace is timed with the vertical rocking of the antenna.

Fig.7-9 - Screw Scanning

On the other hand, the motion of the spots on the screen along the sweep trace to right and left is timed with the rocking of the antenna along the azimuth. In this case, the reflected signal also governs the brightness of the pip, which is at its greatest when the radar is on target.

In order to be subject to monitoring, every point in the search zone must fall within the electromagnetic beam at least on each of two successive rotations of the
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antenna on the azimuth. If the sector covered by the elevation sweep is $b(\theta)$ radian, and the vertical width of the antenna diagram is $\gamma(\theta)$ radian, our condition of double coverage will require two turns of the antenna on the azimuth for observation of a sector. Therefore, when the azimuth angular velocity of the beam is $\omega(\varphi)$ radians per second, and the azimuthal sector of observation is φ radian, the following relationship between the angular velocities of the beam on the azimuth and in elevation must be adhered to:

$$\omega(\varphi) = \frac{b(\theta)\omega(\varphi)}{2\varphi} \quad (7-2)$$

The minimal pulse frequency required by the indicator is related to the horizontal width $b(\varphi)$ of the pattern, and to the azimuthal velocity $\omega(\varphi)$ by eq.(7-1), as in circular scanning.

When employed in ground-based and naval aircraft search stations, circular scanning follows a pattern of 360° rotation along the azimuth ($\varphi = 2\pi$), and a 90° elevation sector. It is also used in aircraft-based sets for searching a sector above and below the aircraft's horizon, and over a 180° sweep in the azimuthal plane.

The main shortcoming of screw scanning is its inadequacy for directly tracking fast-moving targets.

The second appearance of the reflected signal on the indicator screen takes place after about ten seconds have elapsed. This is a period sufficient for the target to be lost during uninterrupted observation of the entire search zone.

Section 7-8. Conical Scanning

If the exciter of a paraboloid be moved relative to the focus, the direction of the electromagnetic beam will vary from the focal axis of the paraboloid. When the exciter is shifted downward, the beam rises and when it is shifted upward the beam dips. When it is moved to the right, the beam is moved to the left, and when the exciter is moved to the left, the beam will deviate to the right from the axis.

of the paraboloid. And if an exciter dipole positioned assymmetrically relative to the focus of a stationary paraboloid is rotated around the focus, the electromagnetic beam will describe a cone. The axis of the cone will coincide with the axis of the paraboloid. This shifting of the beam in space is termed conical scanning (Fig.7-10).

The angle between the generatrices of the cone, and the width of the electro-

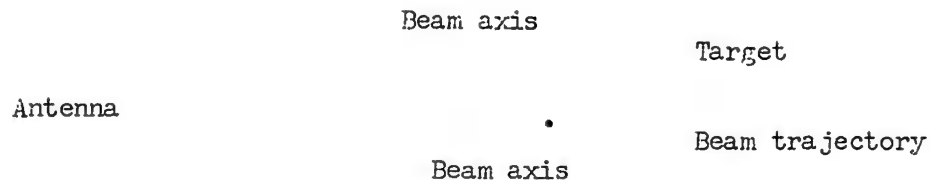


Fig.7-10 - Conical Scanning

magnetic beam, are such that the positions of the rotating lobe of the pattern in any longitudinal section of the cone will overlap. If a single antenna is used for

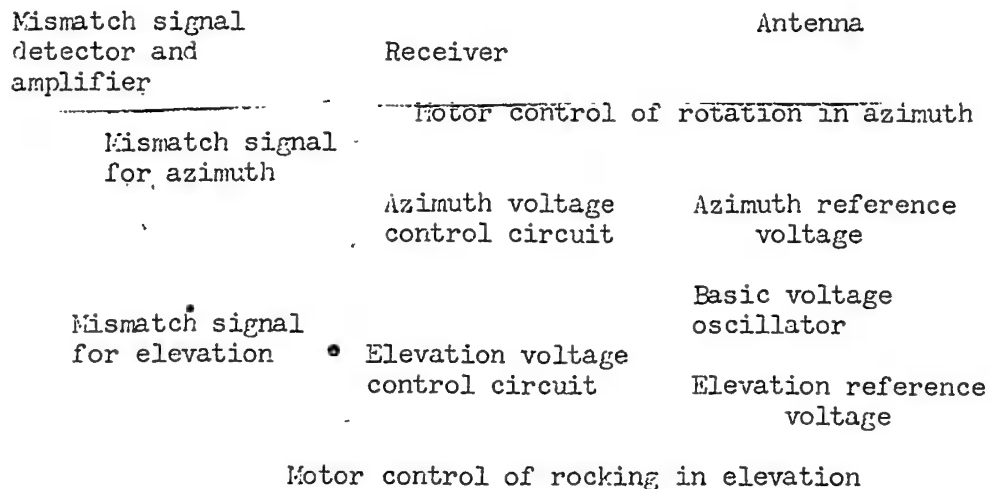


Fig.7-11 - Automatic Tracking Circuit

transmission and reception, then, when the axis of the scanning cone is not in line with the target, the energy of the initial radiation, as well as the energy reflected ^{STAT}

ted from the target, will differ when the beam is in different positions. The reflected signal will vary as the beam rotates. Only in the direction of the axis of the cone will the amplitude of the reflected signal from the target always be the same, regardless of the position of the beam. Therefore, the position of the axis of the paraboloid is easily noted and employed for taking both azimuth and elevation bearings. The bearing is accurate to within a few hundredths of a degree.

It is not difficult to see, by comparing Figs. 7-10 and 7-2, that for purposes of taking bearings, conical scanning is a further development of the equal-strength-of-signal-zone method. Conical scanning is used in the microwave-band sets used for gun-laying with automatic target tracking, and also in certain types of aircraft radars.

Use of conical scanning for automatic tracking in azimuth and tilt planes may readily be understood on the basis of the following example.

If the position of the antenna is such that the axis of the scanning cone does not entirely coincide with the direction of the target, the reflected signal obtained at the receiver output (Fig. 7-11) on rotation of the exciter dipole will be modulated sinusoidally with a frequency equal to the number of revolutions of the dipole per second. At the detector output, a mismatch signal will form from the modulated reflected signal, and two signals of error will then be amplified. The amplitude of the error signal for the azimuth will be greater at a greater deviation of the axis of the cone from the target along the azimuth, while the amplitude of the elevation mismatch signal will be greater at a greater deviation of the axis of the cone from the target in elevation. If the axis of the cone is directly on target, the mismatch signals are zero.

An oscillator to produce sinusoidal reference voltages is mounted on the axis of rotation of the dipole. Its frequency is also equal to the number of rotations of the dipole per second. The circuit for the controlling azimuth voltage as affected simultaneously by the azimuth mismatch signal and the azimuth reference volt-

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age is developed to accord with the sign of the voltage controlling the motor causing the antenna to revolve in the azimuthal plane. Likewise, the circuit for the controlling elevation voltage is affected simultaneously by the elevation mismatch signal and the azimuth reference voltage is developed to accord with the sign of the voltage controlling the motor causing the antenna to rock in a vertical path. The controlling voltage circuits act on the motors so as to cause the antenna always to take a position in which the axis of the cone is virtually coincident with the direction to the target, and the mismatch signals are approximately zero. The azimuth and elevation are read by means of pointers always on target.

In determining the pulse frequency to be used in conical scanning it is necessary to remember that it must always exceed a minimum.

$$F_{u \text{ min}} = n \frac{\omega_{sp}}{b} \quad (7-3)$$

Here b is the width of the antenna radiation pattern, in radians;

ω_{sp} is the angular velocity of rotation of the dipole;

n is the minimum number of pulses needed by the given type of indicator to illuminate the target as long as it is in the zone of the antenna beam.

Section 7-9. Spiral Scanning

~~In stations with automatic tracking, spiral scanning is used in combination~~
with conical scanning, the same antenna being used for both.

If conical scanning is to be used for obtaining bearings and automatic tracking, the preceding spiral scanning is used for very quick location of targets within the portion of a sphere bounded in elevation and azimuth. In spiral scanning, not only does an exciter placed assymmetrically relative to the focus undergo rotation around that focus, but the paraboloid itself rocks in the vertical plane. If the paraboloid undergoes no change in azimuthal position, the trajectory of a point on the circle described by the beam due to the rotation of the dipole, changes into a spiral because of the rocking of the paraboloid. The coils of the spiral lie on

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the surface of a sphere of radius R (Fig.7-12).

The ratio between the rate of rotation of the dipole and the rate at which the paraboloid rocks must be set in such a manner that any point in the space being searched falls within the beam in not less than two successive coils of the spiral.

It can be shown that the required ratio of angular velocities is as follows:

$$\omega_{\text{rock}} = \frac{b\omega_{\text{sp}}}{4\pi} \quad (7-4)$$

in which ω_{rock} is the angular velocity at which the paraboloid rocks in the vertical plane;

ω_{sp} is the angular velocity of rotation of the exciter;

and b is the beam width.

In spiral scanning, the pulse frequency must exceed the following minimum

$$F_{i \text{ min}} = n \frac{\omega_{\text{sp}}}{b} \sin \frac{\beta}{2} \text{ [impulses/second]} \quad (7-5)$$

where n is the minimum number of pulses for the given indicator;

β is the sector through which the paraboloid rocks in the vertical plane.

The prime advantage of spiral scanning over the other types is the reduction

in scanning time to a period of some-

times not more than one second.

Beam trajectory

Figure 7-12 illustrates the trajectory of the electromagnetic beam in spiral scanning when there is no change in the position of the paraboloid in the azimuthal plane. However, a combination of spiral scanning with circular rotation of the antenna in the plane of the azimuth, and determining of bearing by maximum

Fig.7-12 - Spiral Scanning

imum signal strength permits the use of a circular scanning indicator and reading

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the azimuth of the target directly on the screen, as with standard circular scanning. When this is the case, the indicator screen shows an image identical to that obtained in circular scanning (see Fig.3-4b). The elevation of the target is measured only on switching from spiral to conical scanning.

Section 7-10. Search With the Cosecant-Squared Beam

In designing surface spotter radars, it is quite logical to desire that the signals reflected from aircraft at a given altitude, regardless of range, be of approximately identical amplitude. Solution of this problem is possible only if the antenna array employed provides a specific unequal distribution of energy in the flare of the electromagnetic beam.

The inclined ranges R of points at unknown elevation h are proportional to the cosecant of the elevation θ . From Fig.7-13b we find that, in reality:

$$R = \frac{k}{\sin \theta} = h \operatorname{cosec} \theta \quad (7-6)$$

Allowing for the fact that the field intensity declines in proportion to the first power of the range, the following expression may be written for the field intensity of the outgoing beam at a point at altitude h :

$$E_h = \frac{k_g}{R} = \frac{h E}{h \operatorname{cosec} \theta} \quad (7-7)$$

in which k_g is a coefficient of proportionality depending upon the radiated power of the antenna and allowing for its directivity in terms of field intensity.

In order that the amplitude of the signals reflected from aircraft flying at altitude h and at different distances be identical, it is necessary in the first place that the outgoing beam be of identical field intensity at all points on that line. The required distribution of radiation may be attained by means of a special antenna, the field intensity of which varies with the cosecant:

$$E = E_{\max} \operatorname{cosec} \theta$$

where E_{\max} is the field intensity in the direction of maximum radiation.

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From this we obtain:

$$\frac{E}{E_{\max}} = \operatorname{cosec} \theta$$

Therefore, the factor k_E employed in eq.(7-7) will also be proportional to the magnitude of $\operatorname{cosec} \theta$ when this type of antenna is used, i.e., $k_E = k \operatorname{cosec} \theta$, where k is a constant. Taking this into consideration, we obtain the following from eq.(7-7):

$$E_h = \frac{k \operatorname{cosec} \theta}{h \operatorname{cosec} \theta} = \frac{k}{h} = \text{const}$$

Thus, when a cosecant-pattern antenna is used, the field intensity of the outgoing beam does not depend upon the elevation and is constant at all points in a

Fig.7-13 - Search With the Cosecant-Squared Beam

line of equal altitude.

The return paths from points on a line of equal altitude are naturally in agreement with the same expression (7-6). Examination of the cosecant shape of the pattern when the antenna is used for reception readily shows that the signal reflected from a given target at all points on a course of equal altitude and regardless of range R will reach the receiver input at the same amplitude.

Thus, solution of this problem is provided by a directional antenna, the field intensity of which is proportional to the cosecant of the elevation (Fig.7-13c), and whose radiated power is proportional to the square of the cosecant of the elevation. STAT

tion. This type of antenna is called a cosecant-squared antenna. It radiates the larger part of its energy parallel to the surface of the earth, and the smaller part

Inclined beam

Vertical beam

Horizon line

Fig. 7-14 - V-beam Antenna Pattern

at angles thereto. The antenna reflector in this case is made of two parabolic parts of different focal lengths, the lower portion being thrust somewhat forward (Fig. 7-13a).

A cosecant-squared beam may be obtained not only by the proper design of reflector components, but also by the use of a complex exciter of special design. This exciter has to provide the dephasing of the energy radiated by various components at the surface of the normal parabolic reflector, as required

Fig. 7-15 - Use of a V-beam Pattern to Determine the Elevation

- a) Moment of spotting by vertical beam;
- b) Moment of spotting by inclined beam

Section 7-11. V-beam Search

For search of surrounding space and measurement of coordinates in all three

dimensions, a special transmit-receive antenna with a V pattern is highly useful.

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This antenna consists of two rigidly attached parts rotating jointly in the azimuthal plane. The beams of both parts of the antenna are quite narrow in the horizontal plane and quite broad in the vertical. The beam of one part of the antenna is always in the vertical position, and the beam of the other is at 45° to the horizon. As a result, the resultant beam of the antenna is in the shape of the Latin letter V (Fig. 7-14).

In taking a bearing by determination of maximum signal strength, each target is spotted twice as the antenna rotates: first by the vertical beam and then by the inclined beam.

Spotting the target with the vertical beam makes it possible to measure its range on the incline R and its azimuth φ . This moment of discovery is shown in Fig. 7-15a. The triangle OKZ gives the elevation of the target:

$$\theta = \sinh \frac{h}{D} \quad (7-8)$$

where h is the altitude of the target;

D = OK is the horizontal range (projection of the inclined range onto the horizontal plane).

The moment at which the inclined beam then spots the target corresponds to the measurement of another azimuthal magnitude $\varphi + \Delta\varphi$, where $\Delta\varphi$ is the angle of rotation of the antenna between the moments at which the target is spotted by the vertical and horizontal beams. Due to the fact that the axial plane of the inclined slope forms an angle of 45° with the earth, we find the following from the triangle Z'OK (Fig. 7-15b):

$$\Delta\varphi = \sinh \frac{h}{D} \quad (7-9)$$

Comparing eqs. (7-8) and (7-9), we find that

$$\theta = \Delta\varphi$$

Thus, taking an elevation bearing by means of a V-beam antenna comes down to reading the difference between the azimuths of the target at the instants when it is

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spotted by the vertical and horizontal beams of the antenna.

Section 7-12. Advantages and Disadvantages of Pencil-Beam Antennas for Search and Bearings

The degree of directivity of receiver antennas is determined, in radiolocation, primarily by the need for precision in the measurement of the angular coordinates of a target in direction finding.

Higher directivity is also quite useful in increasing the effective radius of an antenna. The sharper the directivity of a transmitting antenna, the more strongly it concentrates electromagnetic radiation in the direction of the target, and the stronger is the field of primary radiation of the target. On the other hand, like an optical lens which focuses incident light rays of a single point, a directional receiving antenna collects the energy of the electromagnetic rays reflected by the target at the moment it is spotted, and thereby amplifies the e.m.f. emitted by the antenna. Thus, a most important problem in the design of radar stations is that of increasing the directivity of the antennas.

On the other hand, extreme narrowing of the directional beam makes it difficult to find targets. The narrower the electromagnetic beam of the transmitting antenna, the smaller the number of targets it is capable of revealing simultaneously, particularly at short ranges. Narrowing of the beam makes it necessary to increase the speed with which it is swept, and this often creates serious engineering problems. To obtain a very highly directional beam is difficult in itself, as narrowing of the pattern is attained by making the antenna still more complicated and by increasing its linear dimensions relative to the operating wavelength.

Radiation of the target with the direct pulse is possible only during the time interval in which the electromagnetic wave moving through space follows the target with its full width. To increase the probability of spotting when scanning with an electromagnetic beam at high speed, it is necessary either to considerably increase

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the pulse frequency, which, as we have seen, limits the range of the station, or to increase the pulse power of the transmitter.

Thus, the advantages provided by pencil-beam antennas are high accuracy of the bearings obtained, and a clear improvement in the range of a station of given power, while the disadvantages lie in the difficulty of finding the target and the relatively larger linear dimensions of the antenna. The decision as to type of antenna radiation pattern and array is determined by the tactical and technical specifications for the particular locator station: military purpose, radius of effectiveness and operating wavelength.

CHAPTER X

RADAR RECEIVERS

Section 10.1 Purpose, and Specifications to be Met

The purpose of a radar receiver is to amplify the radio echo pulses received by the antenna, to eliminate noises, and to convert the radio pulses into video pulses, i.e., DC pulses.

The receiver unavoidably amplifies not only the useful signal but the chaotic interfering voltages called noises. The echo signal will not be visible on the indicator screen if the signal, amplitude is lower than the noise level. Therefore, the first and most important specification for a receiver is that it produce minimum set noise.

The problem of receiving the echo signal is complicated as the range to target increases, in view of the marked weakening of the amplitude. The power of the signals emitted by very distant targets does not exceed a fraction of a micromicro-watt, corresponding to an antenna emf of the order of a few microvolts. Normal operation of indicators requires the application of power of the order of one watt, corresponding to an output voltage of several volts or even several tens of volts. Therefore, in addition to producing minimal set noise, a radar receiver must be highly sensitive and have very great gain, of the order of several millionfold. This is the second requirement.

In order not to introduce additional errors into the range measurements, the receiver must ensure minimum distortion of the echo-pulse envelope and exceedingly

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0 brief transit times from input to output. This is the third requirement.

2 For certainty of reception of echo signals, and to attain high gain, the re-
4 ceiver must tune easily to the operating frequency of the transmitter, ensuring
6 maximum sensitivity there, and must readily follow any fluctuations in the trans-
8 mitter frequency. This is the fourth requirement.

10 Despite marked attenuation in the transmit-receive switch, the energy of a
12 powerful transmitter will nevertheless leak into the receiver to some degree. There-
14 fore, the receiver circuit must contain its own protection from overloads by the
16 transmitter signal, and also from the effects of powerful outside interference.
18 This is the fifth requirement.

20 All other requirements to be met by radar receivers are dictated by these
22 five major requirements and derive directly from them.

24 Section 10-2. Receiver Set Noises

26 The theory of this most important question has been worked out in detail in
28 the works of V.I.Siforov and A.A.Kolosov, and also by V.A.Kotel'nikov and others.

30 Set noises arise from phenomena of accidental and random character and giving
32 rise to chaotic, fluctuating voltages and currents in the receiver circuits.

34 Fluctuation voltages and currents show the most extreme variations in frequency and
36 phase, due to their accidental nature. Noises may therefore be represented by a
38 continuous series of a very large number of harmonics, always embracing the entire
40 receiver bandwidth, regardless of how great it may be, and regardless of the fre-
42 quency to which the receiver is tuned. Noise energy consists of the total energies
44 of all the noise harmonics and therefore increases in proportion to the widening
46 of the receiver bandwidth.

48 Oscillating-Circuit Noises

50 The electrons in any conductor are in constant random thermal agitation. The
52 electron in random agitation may be regarded as an elementary random current

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creating an elementary voltage against the resistance of the conductor even when no external source of electricity is present. The resultant noise voltage is composed of the totality of such elementary random voltages. The total noise power in a circuit is greater, the higher the temperature and the wider the pass-band of the circuit.

The noise power in a circuit is:

$$P_n = 4kT\Delta F = \frac{U_n^2}{R_e} \quad (10-1)$$

where T is the absolute temperature in degrees Kelvin;

$k = 1.38 \times 10^{-23}$ joules per degree denotes a scale factor;

ΔF is the pass-band, in cycles;

R_e is the active component of the equivalent resistance of the circuit, in ohms.

From this we obtain the square of the noise voltage, in volts

$$U_n^2 = 4kTR_e\Delta F \quad (10-2)$$

or

$$U_n = 2 \sqrt{kTR_e\Delta F}$$

If R_e is expressed in k-ohms and ΔF in kilocycles, we will have, at room temperature ($T = 290^\circ K$)

$$U_n \text{ (mkv)} = \frac{1}{8} \sqrt{R_e \Delta F} \quad (10-3)$$

Thus, if the pass-band is $\Delta F = 1$ megacycle and $R_e = 1000$ ohms, the noise voltage will be

$$U_n = \frac{1}{8} \sqrt{1 \times 10^3} = \frac{31.6}{8} \approx 4 \text{ mkv}$$

The noise voltage applied to the antenna, due to thermal agitation of the charged particles of the environment, may be calculated by an equation analogous to eq.(10-2):

$$U_{nA}^2 = 4kT_A R_2 \Delta F$$

where T_A is average antenna temperature; and

R_2 is the radiation resistance of the antenna.

Receiver-Tube Noises

Tube noises are determined primarily by the shot effect due to the random thermal inconstancy of cathode emission. The shot effect causes a fluctuation of the plate current which, in turn, results in a fluctuation voltage U_{na} in the plate load resistance.

The space charge in the tube, absorbing the shocks from the electron stream moving from cathode to plate, modifies the fluctuation and reduces the noise. On the other hand, the presence of positive ions in the tube, due to inadequate hardness of the vacuum tends to increase noise. This may be explained by the fact that the positive ions settling on the cathode and becoming neutralized there, complicate the escape of electrons and impair the emission. As a result of the quantitative fluctuation in emission, the neutralizing positive ions produce further fluctuations in the emission currents and increase the noise.

Fluctuations in the plate current increase with its level. As a result, a larger number of electrons undergo fluctuation. The existence of positively-charged grids absorbing part of the electron stream moving toward the plate, increases the noise, since the inconstancy of the emission current aggravates the inconstancy of its distribution between the electrodes. Pentodes make more noise than triodes.

The fluctuation noise voltages U_{na} , discriminated across the resistance of

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the plate load, may be replaced by an equivalent noise voltage U_{nc} cut into the grid circuit. Obviously

$$U_{nc} = \frac{U_{na}}{K}$$

where K is the gain of the stage using the given tube in the given operation.

This substitution makes it possible to regard the tube as noiseless but as having in its grid circuit a noise-voltage oscillator U_{nc} (Fig.10-1). The most convenient method is to substitute the voltage source U_{nc} by an equivalent noise resistance, R_n . The noise voltage across this resistance can be calculated in the usual manner (by eqs.10-2 or 10-3) and taken as the U_{nc} voltage.

- Two equations for the noise resistance of an amplifier tube are known:

for triodes

$$R_n \text{ (kilohms)} = \frac{2.5 - 3}{S \text{ (ma/v)}} \quad (10-4)$$

for pentodes

$$R_n \text{ (kilohms)} = \frac{I_a}{I_a + I_{(g)}} \left[\frac{2.5}{S} + 30 \frac{I_{(g)}}{S^2} \right] \quad (10-5)$$

where I_a and $I_{(g)}$ are, respectively, the plate and screen currents, in milliamperes.

As indicated in eq.(10-4), the noise resistance and the noise both increase as the transconductance declines, since a decline in transconductance is equivalent to a reduction in the dynamic gain of the cascade K . When the tube is used as a converter the noise resistance increases 3 to 5-fold, since the conversion transconductance is less than the transconductance of the same tube when used as an amplifier.

The considerable time required for the electron transit increases the noise, since the fluctuations in the electron stream are supplemented by fluctuations in the currents applied in the electron circuits. Tubes make more noise in the ultra-short-wave band. Noise reduction requires a reduction in the electron transit

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time.

Section 10-3. Noise Figure and Sensitivity

It is convenient to calculate the noises in all stages of the receiver, with allowance for the gain of these stages, in terms of the grid circuit of the first

R_n

U_{nc}

tube. The power of the total (discriminated) noises $R_{\Sigma n}$ in the grid circuit of the first tube is composed of the powers of the lumped noises in the following receiver components: P_{nA} in the antenna, P_n in the input circuit, P_{nt} of the first tube, $P_{n\ tbs}$ for all the other tubes, i.e.,

Fig. 10-1 - Noise-Voltage Oscil-

lator $U_{nc} = \frac{U_{nA}}{K}$ in the Tube

Grid Circuit, used as Voltage

across the Resistance R_n

$$P_{\Sigma n} = P_{nA} + P_{n\ in} + P_{nt} + P_{n\ tbs}$$

Considering that the power of the tubes is proportional to the square of the corres-

ponding noise voltages, we may write:

$$U_{\Sigma n}^2 = U_{nA}^2 + U_{n\ in}^2 + U_{nt}^2 + U_{n\ tbs}^2$$

From this we find that the total noise voltage effective in the grid circuit of the first tube, is equal to:

$$U_{\Sigma n} = \sqrt{U_{nA}^2 + U_{n\ in}^2 + U_{nt}^2 + U_{n\ tbs}^2}$$

The receiver noises have come to be characterized by the noise figure (noise factor), which is usually determined by means of a signal generator.

The noise figure gives data on the factor by which the power of the signal P_{cA} and of the noises P_{nA} delivered to the receiver by the signal generator, are greater than the signal-to-noise ratio, $P_{c\ out}/P_{n\ out}$ at the receiver detector

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input. Thus, the noise factor is

$$N = \frac{\frac{P_{cA}}{P_{nA}}}{\frac{P_{c out}}{P_{n out}}} \quad (10-6)$$

In determining the noise factor, the internal impedance of the signal generator, substituting for the antenna, must be equal to the input impedance of the receiver.

From the determination of the noise factor in eq.(10-6) it follows that the power of the signal reaching the receiver input from the antenna is

$$P_{cA} = P_{nA} N \left(\frac{P_{c out}}{P_{n out}} \right) \quad (10-7)$$

The signal-to-noise ratio at the receiver detector input, necessary for satisfactory reception of signals, i.e., the quantity $D = \frac{P_{c out}}{P_{n out}}$, will be denoted below as the discrimination factor.

By analogy to eq.(10-1), the antenna noise power is:

$$P_{nA} = 4kT_A \Delta F_{pb} \quad (10-8)$$

Consequently, the required minimum signal power at the receiver input will be

$$P_{pb min} = 4kT_A \Delta F_{pb} N D \quad (10-9)$$

This power is defined as the sensitivity of a radar receiver.

Equation (10-9) shows that, at a given pass band ΔF_{pb} and a known discrimination factor D , the sensitivity of the receiver is governed by its noise factor N .

The magnitude of D depends upon the shape, length, and frequency of the pulse and also on the indicator and receiver parameters. The smallest possible discrimination factor is obtained when D is unity.

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Section 10-4. Input Impedance of a Tube in Ultrashort-Wave Operation

Among the various works of Soviet scientists that have played a pioneering role in the development of ultrashort-wave tube engineering, particular credit is due to the work of Prof. V.I.Siforov who, working on the basis of the general theory of active linear electric networks, developed by E.V.Zelyakh, V.I.Kovalenkov, and V.I.Siforov, involved an orderly theory of amplifier tube function in the ultrashort-wave band.

In accordance with this theory, the current within an electron tube is, in the general case, not equal to that in the corresponding external circuit.

Two types of current must be distinguished for an electron tube:

- a) Conduction current i_k , serving to propel the electric charges in the space between the various tube electrodes; and
- b) Displacement current or capacitance current i_c , generated by a change in the voltage on the electrodes, i.e., by the presence of a varying electric field concentrated in the capacitance between the electrodes.

The total current within the tube is equal to the sum of the conduction current and the displacement current.

The electrons moving in the space between the tube electrodes create their own electric field. Under the influence of this field, electric charges are induced at the surfaces of the electrodes, while in the wire connecting the electrodes between which the electrons flow, current is induced regardless of whether the electrodes do or do not intercept a portion of the electron stream. If the magnitude or velocity of the moving charge varies with time, the induced current also changes.

A displacement current also arises between the electrodes when they are under alternating voltage. Therefore, the total current in the wire connecting the electrodes is equal to the sum of the induced and displacement currents. Thus, the appearance of induced currents in the external circuits of tube electrodes is due

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to the inductive effects of the electron flow within the tube.

Specifically, the electrons passing through the grid turns on their way from cathode to plate, induce a positive charge in the grid. This charge increases as the electrons approach the grid and declines as they retreat into the distance. An increase in the positive charge induced is equivalent to the appearance of an induced grid current i'_g , while a decline in this charge is equivalent to the appearance of an induced current i''_g . These currents are opposite in direction (Fig. 10-2). Therefore, the resultant induced current in the external grid circuit is

$$i_F = i'_g - i''_g$$

The magnitude of the induced current i_g is significantly dependent on the electron transit time from cathode to plate.

With long and short waves, the transit time of electrons is negligible by comparison with the duration of the oscillating voltages acting on the tube electrodes. It may be considered that, in the given case, the phase of the alternating grid voltage does not have a chance to change within the transit time. The velocity of the electrons approaching and receding the grid will be determined by the same phase of grid voltage. The increase in the positive charge on the grid, induced by the approaching electrons, is compensated by its decrease under the influence of the departing electrons. The currents i'_g and i''_g will thus be equal and no resultant current will be induced in the grid circuit. The flow of electrons will show virtually no inductive effect in the long- and short-wave bands.

The electron cathode-to-plate transit time, which is of the order of 10^{-9} sec in ordinary tubes, becomes equal to the period of oscillation, in the ultrashort-wave band. Thus, at a 30-cm wavelength, the period of oscillation T is 10^{-9} sec. In any case, on ultrashort-wave, during the electron transit time, constituting a considerable proportion 2π of the oscillation period, the phase of the alternating voltage on the grid is able to change by an angle φ_T . Therefore, the velocity of

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the electrons passing through the grid and that of the electrons that have just been emitted by the cathode, will be determined by different grid-voltage instan-

taneous values and will not be equal. The induced currents i'_g and i''_g will differ. The resultant induced current i_g appears in the external grid circuit. Its first harmonic, i.e., the component of the frequency of the alternating voltage on the grid, will be considerable.

Fig.10-2 - Currents Induced in the Grid Circuit due to the Inductive Effects of the Electron Flow:

i'_g , the current Induced by Electrons Approaching the Grid;

i''_g , Current Induced by Electrons

Moving Away from the Grid

Thus, the effect of transit time due to inductive action of the electron stream in the ultrashort-wave band is a source of considerable grid currents, even when the negative displacement on the grid is such that grid currents would be incapable of appearing in the long and short waves.

In order to investigate the question of input impedance, let us consider the pentode as the most important type of amplifier tube.

The input impedance of a tube is the ratio of the total input voltage \bar{U}_{in} to the total input current \bar{I}_{in} , i.e.,

$$\bar{Z}_{in} = \frac{U_{in}}{I_{in}} \quad (10-10)$$

Let us examine Fig.10-3a. The applied current required by the pentode from the source of high-frequency voltage branches in part at the cathode through the interelectrode capacitance (current \bar{I}_{gf}) and in part through the screen grid circuit across the capacitance $C_{(g)g}$ (current $\bar{I}_{(g)g}$). Therefore,

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$$\bar{Z}_{in} = \frac{U_{in}}{I_{gf} + I_{(g)g}} \quad (10-11)$$

The quantities entering here are complex, since the presence of reactance parameters will cause the currents \bar{I}_{gf} and $\bar{I}_{(g)g}$ to be out of phase with the voltage

Fig.10-3 - Circuits and Vector Diagram of Pentode Used

in the Ultrashort-Wave Band

at the input. In addition, these currents would be purely capacitive if it were not for the effect of the electron transit time and the influence of the active leaks of interelectrode capacitance. Inductivity in the tube leads would be lacking. The current \bar{I}_f , at inductance L_f , creates a voltage \bar{U}_{Lf} , which leads the current by 90° . The total input voltage \bar{U}_{in} is equal to the geometric sum of the voltage between grid and cathode, and the voltage on the inductance of the cathode lead. The current \bar{I}_{gf} , branching through the capacitance C_{gf} at the cathode, is governed by the voltage \bar{U}_{gf} between the control grid and the cathode, and leads that voltage by 90° , being in phase with the voltage \bar{U}_{Lf} . The capacitance $C_{(g)g}$ is inserted between the control grid and the ground. Therefore, the current $\bar{I}_{(g)g}$ is governed by the total input voltage \bar{U}_{in} and leads it by 90° . The total input current \bar{I}_{in} is equal to the geometric sum of the currents \bar{I}_{gf} and $\bar{I}_{(g)g}$. It leads the input voltage \bar{U}_{in} by less than $\frac{\pi}{2}$. This means that the input impedance of

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the pentode is not purely capacitive but complex in nature.

The complex input impedance of the pentode may be conceived of as a parallel connection of the input capacitance C_{in} and the resistance R_{in} . Excluding the effect of electron transit time, the resistance R_{in} is equal to the resistance R_{inL} , governed by the effect of the inductance L_f of the cathode lead:

$$\left. \begin{aligned} C_{in} &= C_{gf} + C_{(g)g} \\ R_{inL} &= \frac{1}{\omega^2 C_{gf} L_f S_f} \end{aligned} \right\} \quad (10-12)$$

where ω is the angular frequency of the input voltages; and

S_f is the cathode current transconductance.

Due to the inductive effect of electron flow in the circuits of the control and screen grids, induction currents are generated which govern the active input impedance:

$$R_{in\tau} = \frac{1}{k S \omega^2 \tau_g^2} \quad (10-13)$$

where k is a factor whose value is approximately $\frac{1}{20}$ for flat-electrode tubes;

τ_g is the transit time of the electrons between cathode and the control grid; and

S is the transconductance of the total current from cathode to plate and to screen grid.

The resistance is inserted parallel to the input capacitance and, consequently, parallel to the resistance R_{inL} (Fig. 10-3c). Therefore, the total active input resistance can be found from the equation:

$$\frac{1}{R_{in}} = \frac{1}{R_{inL}} + \frac{1}{R_{in\tau}} = \omega^2 C_{gf} L_f S_f + k S \omega^2 \tau_g^2$$

or

$$R_{in} = \frac{1}{\omega^2 (C_{gf} L_f S_f + k S \tau_g^2)} \quad (10-14)$$

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Two conclusions can be drawn from eq.(10-14):

a) In order to increase the input impedance, the tube must be so designed as to provide minimum inductance of the cathode input and minimum electron transit time:

b) The resistance of a given tube is inversely proportional to the square of the voltage frequency on the control grid.

We may, therefore, write

$$R_{in} \text{ (megohms)} = \frac{k}{f^2 \text{ (megacycles)}} \quad (10-15)$$

The magnitudes of the K factor for the most important types of amplifier tubes are shown in Table 10-1.

Table 10-1

| Type of Tube | K, megohms/mc ² | Type of Tube | K, megohms/mc ² |
|--------------|----------------------------|--------------|----------------------------|
| 6P9 | 3.0 | 2K2M | 65 |
| 6P3S | 6.5 | 6SLZh | 160 |
| 6Zh4 | 7 | 6SLP | 160 |
| 6Zh8 | 20 | 6Zh1Zh | 200 |
| 6Zh3P | 37 | | |

Section 10-5. A Tube for an Ultrashort-Wave Amplifier

An increase in transit time, combined with a reduction in wavelength results in a reduction in R_{in} .

An important consequence of the reduction in input impedance is an increase in the power required by the grid circuit:

$$P_g = \frac{U_g^2}{R_{in}} \quad (10-16)$$

At a given voltage U_g , the power of the grid circuit increases with decreasing R_{in} .

input impedance as the wavelength decreases. The power gain of the cascade then is

$$K_p = \frac{P_{out}}{P_{in}} \quad (10-17)$$

decreasing in the ultrashort-wave band with decreasing wavelength, due to a rise in P_{in} even when the power at the output P_{out} is stable.

Thus, in the ultrashort-wave band, the tube itself consumes a relatively large portion of its own capacity for amplification.

In addition to a worsening of the amplifying properties of the tube, the gain of a cascade is further reduced in the ultrashort-wave range since the low input impedance as well as the input capacitance of the tube in the next stage by-pass the plate charge of the preceding stage to such an extent as to make its real level quite small. Along with the reduction in resistance of the plate charge when the internal impedance of the tube is high, the output power declines considerably.

Thus, in the ultrashort-wave band, amplification encounters major difficulties due to the effect of the electron transit time and the influence of the inductances of the tube leads and the interelectrode capacitances.

a)

b)

Fig.10-4 - Ultrashort-Wave Receiver Tubes, Including Low-Power Amplifiers

a - 6Zh1Zh acorn pentode; b - 6K1P bantam tube

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One measure to facilitate an increase in input impedance is a provision in the design, for low inductance at the cathode input. This is attained by shortening the cathode pins, enlarging their diameter, and increasing their spacing. In certain types of amplifier tubes for the ultrashort-wave band, the cathode lead takes the form of several (usually four) pins in parallel. Similarly, the inductance of the plate lead is reduced because the plate is also provided with a lead by means of two spaced conductors.

The reduction in interelectrode capacitances can be obtained either by reducing the surface areas of the electrodes or by increasing their spacing. Small electrodes compel the tube to work under more severe thermal conditions, since the amount of heat the electrodes are to dissipate declines. In the long run, this also limits the oscillating output power. However, ultrashort-wave receiving tubes, including low-power amplifiers, usually operate as voltage rather than as power amplifiers. Therefore, the small dimensions of the electrodes in this case are entirely in order and constitute a standard design characteristic of ultrashort-wave receiving tubes, including low-power amplifiers. As far as increasing the electrode spacing is concerned, this procedure causes an increase in the electron transit time and is not an acceptable method when the feed voltages at the electrodes are low.

Figure 10-4 shows the most characteristic designs of ultrashort-wave receiver tubes.

Section 10-6. Special Features of Radar Receivers

The specific features of radar receivers are determined by the fact that they work in pulses and operate in the radar band of working waves.

The first group of special features of radar receivers is governed by the very wide band required to pass pulse signals reflected from a target, i.e., radio pulses (before detection) and video pulses (after detection). Let us explain this

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in terms of an ideal rectangular pulse.

In accordance with Fourier's theorem, well-known in electrical engineering, a periodically repeated video pulse voltage, like all periodic voltages of nonsinusoidal shape, can be considered as the result of the superimposition of an infinite

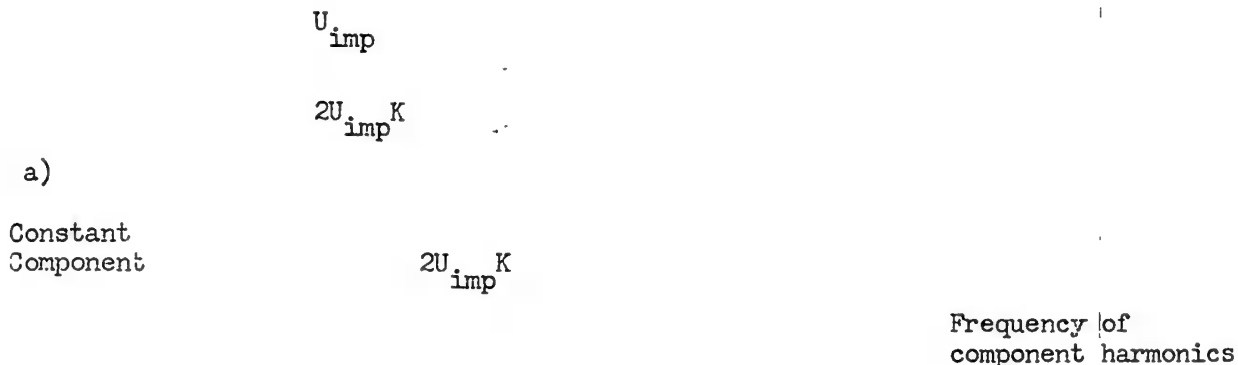


Fig.10-5 - Spectrum of Periodically Repeated Rectangular
Video Pulse

a - Amplitude; b - Energy

series of constantly acting components.

When the starting point of the time scale has been properly selected, as illustrated in Fig.10-5, the components of a rectangular video pulse will include constant cosinusoidal components, invariable in magnitude and polarity, called harmonics. The frequency of the harmonics exceeds, by a factor that is an integer n , the pulse repetition rate F , where n is the number of harmonics.

A mathematical analysis of the case presented can be given in the form of the following Fourier series:

$$u(t) = \frac{U_{m \ n=0}}{2} + \sum_{n=1}^{n=\infty} U_{mn} \cos n(2\pi F_u) t$$

or, in developed form,

$$u(t) = \frac{U_{m \ n=0}}{2} + U_{m \ 1} \cos(2\pi F_u) t + U_{m \ 2} \cos 2(2\pi F_u) t + \\ + U_{m \ 3} \cos 3(2\pi F_u) t + \dots + U_{mn} \cos n(2\pi F_u) t$$

The relationship between the instantaneous values of the harmonic components and their phases is such that, beyond the limits within which the video pulse can be sustained, the total voltage of all the components equals zero.

The amplitudes of the harmonics with various order numbers are not identical, although they are proportional to a single magnitude, the pulse value U_{imp} of the voltage. Thus, a doubling of U_{imp} causes the amplitudes of all the harmonics to double simultaneously. The amplitude of the harmonic depends on its order number, the length of the video pulse, τ_u , and the repetition rate F_u .

The following is the theoretical general equation for the amplitude of the harmonic of order number n , for a video pulse of ideal rectangular shape:

$$U_{mn} = 2U_{imp} k \frac{\sin(n\pi k)}{(n\pi k)} \quad (10-18)$$

where $k = \frac{\tau_u}{T_u} = \tau_u F_u$ is the duty cycle.

The direct component of the video pulse is equal to half the amplitude of a harmonic whose order number is zero ($n = 0$), i.e., of a harmonic whose frequency is equal to zero. We know from the theory of limits that the ratio of the sine of an angle to the magnitude of that angle tends toward unity if the angle tends toward zero.

Within the limits of our case ($n = 0$) we obtain:

$$U_{m \ n=0} = 2U_{imp} k$$

As a result, the direct component of a periodically repeated video pulse becomes

$$U_o = \frac{2 U_{imp} k}{2}$$

or

$$U_o = U_{imp} k \quad (10-19)$$

Equation (10-18) shows that, depending on the number of the harmonic, its amplitude may be either positive or negative, since the amplitude is governed by the function:

$$\frac{\sin (n \pi k)}{(n \pi k)}$$

The amplitude spectrum of the voltage of a periodically repeated video pulse is shown in Fig.10-5a. The adjacent harmonics of the spectrum are separated in frequency by F_u .

The amplitudes of certain frequencies in the spectrum are zero (i.e., certain harmonics do not exist). These are called zero frequencies. In order for the amplitude of a harmonic to equal zero, we must have $\sin (n \pi k) = 0$. This is the case when $n \pi k = \pi, 2 \pi, 3 \pi \dots$, etc., which corresponds to the condition $n = 1, 2, 3$, etc. From this we find the numbers of the harmonics of zero amplitude:

$$n_o = \frac{1}{k}, \frac{2}{k}, \frac{3}{k} \dots,$$

where

$$k = \frac{\tau_u}{T_u}$$

Consequently, the zero frequencies of the spectrum are:

$$\left. \begin{aligned} f_{01} &= F_u n_{01} = \frac{F_u T_u}{\tau_u} = \frac{1}{\tau_u} \\ f_{02} &= F_u n_{02} = \frac{2}{\tau_u} \end{aligned} \right\} \quad (10-20)$$

$$f_{03} = F_{u^{n03}} = \frac{3}{\tau_u} \quad]$$

The same may be said of the current spectrum of a repeated rectangular video pulse.

Each harmonic is of an energy proportional - as we know from electrical engineering - to the square of its amplitude. Therefore, the energy distribution of the harmonics through the spectrum of the video pulse (Fig.10-5b) is more clearly defined than the amplitude distribution.

Now let us proceed to the spectrum of ideal rectangular, periodic radio pulses obtained by modulating high-frequency oscillations with rectangular video pulses.

Modulation by video pulses is a special case of amplitude modulation, in which changes occur only in the amplitude of the oscillations, where they rise from zero to the pulse value and then decline again to zero.

We know from radio engineering that, in sine wave amplitude modification by several frequencies at once, the two sidebands - an upper and lower, each separated from the carrier by the magnitude of the modulating frequency - of each of the modulating frequencies are added to the spectrum of the amplitude-modulated oscillations.

The total frequency band of the amplitude-modulated oscillations is equal to the total width of the upper and lower sidebands, i.e., to a doubling of the highest modulating frequency.

Video pulse modulation is equivalent to simultaneous amplitude modulation by all harmonic components of the video pulse spectrum, each of these components adding its upper and lower sidebands to the radio-pulse spectrum (Fig.10-6). Therefore, regardless of the harmonic number, the frequency band occupied by the radio pulse is twice as wide as the frequency band occupied by the modulating video pulse.

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A comparison of Figs. 10-5b and 10-6 shows it to be quite obvious that, in the video pulse spectrum, the very lowest frequency is equal to the repetition rate F_i , is located in the low (audio) frequency band, and also has a direct component. However, in the radio pulse spectrum, as in that of any HF amplitude-modulated oscillation, low frequencies are entirely absent.

In order for periodic pulses to pass through an electric circuit without significant distortions, it is sufficient to transmit only that portion of their spectrum that is highest in energy. We see from Figs. 10-5b and 10-6 that the maximum energy is found in the frequency band limited (as far as periodic ideal rectangular video pulses are concerned) by the first zero frequency, which is

$$\Delta F_v = \frac{1}{\tau_u} \quad (10-21)$$

while in the case of periodic ideal rectangular radio pulses, the maximum energy is located in the frequency band limited by twice the first zero frequency, or

$$\Delta F_p = \frac{2}{\tau_u} \quad (10-22)$$

where τ_u is the pulse length.

Equations (10-21) and (10-22) show that the pass band of the electric circuit must be wider, the shorter the pulse.

In deciding on the pass band of a radar receiver, the prime consideration is the military purpose of the station it is servicing and the requirements to be met.

In all cases where it is desired to pick up all possible signals, no matter how weak (as in distant-warning or circular-scanning stations), the band width of the receiver in the high and intermediate frequencies is selected so as to provide maximum signal-to-noise ratio, which gives a maximum receiver response. As already discussed in Section 10-2, noises increase in power in direct proportion to the band width. However, starting at a given band width, the signal power increases much more slowly than the noise power. This makes it necessary to find an optimum band,

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providing maximum signal-to-noise ratio. Therefore, significant distortions in pulse shape cannot be avoided, and the band width for the high and intermediate frequencies must be selected at less than that determined by eq.(10-22).

Fig.10-6 - Energy Spectrum of a Periodic Rectangular
Radio Pulse

As shown in practice, the best signal-to-noise ratio exists in the radio receiver of high and intermediate-frequency bandwidth described by:

$$\Delta F_{pn} = \frac{1}{\tau_u} \quad (10-23)$$

If the total maximum expected frequency drift of the transmitter and local oscillator is represented by a quantity Δf_x , a stable reception of signals will require the bandwidth to be expanded to

$$\Delta F_{pn} = \frac{1}{\tau_u} + \Delta f_x \quad (10-24)$$

When automatic local oscillator frequency trim is provided, the magnitude of Δf_x is set at twice the error of the automatic trim.

The bandwidth of the video frequency cascades in the receiver is set at one-half the width of the high-frequency and intermediate-frequency cascades which, in

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0 distant-warning stations, is

$$\Delta F_{\text{video freq}} = \frac{1}{2\tau_u} \quad (10-25)$$

and has been obtained from eq.(10-23).

If the pulse beat is in microseconds, the bandwidth is found directly in megacycles. Thus, if the length of the pulse τ_u is 1 microsecond, the receiver of a distant-warning station equipped with automatic frequency trim with an error of 0.25 megacycles must have a pass band in HF and IF of

$$\Delta F_{\text{pn}} = \frac{1}{1} + 2 \times 0.25 = 1.5 \text{ mc}$$

while, in the video band, it should be

$$\Delta F_{\text{vid freq.}} = \frac{1}{2} = 0.5 \text{ mc}$$

Where high-accuracy stations are concerned (such as a set used for gun-laying), the pass band must be determined from the conditions required for obtaining the desired length of the leading front of the pulse τ_f , with allowance for the frequency drift Δf_x of the transmitter and the local oscillator. The usual solution in this situation is to decide on an HF and IF pass band for the set in accordance with the following

$$\Delta F_{\text{pn}} = \frac{0.7 - 0.9}{\tau_f} + \Delta f_x \quad (10-26)$$

Correspondingly, the pass band in the video frequency must be

$$\Delta F_{\text{video freq.}} = \frac{0.35 - 0.45}{\tau_f} \quad (10-27)$$

The smaller the τ_f , the larger must be the coefficient in eqs.(10-26) and (10-27). For example, in the case of pulses in which the front length τ_f is 0.2 microsecond, and the error of automatic frequency trim is 0.25 mc, the receiver of a gun-laying set must have a pass band for HF and IF, according to

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eq.(10-26), of

$$\Delta F_{pn} = \frac{0.9}{0.2} + 2 \times 0.25 = 5 \text{ mc}$$

When the front length is shorter, the band required is even greater.

Thus, the first group of special features of radar receivers results in the requirement of a very broad pass band, exceeding tens of megacycles in high-accuracy stations. The need to provide a wide pass band also affects the choice of the intermediate frequency (see Section 10-11), which, as distinct from the situation with receivers for radio broadcasting ($F_{np} = 465 \text{ kc}$) is taken as 15 to 100 mc.

The second group of special features of radar receivers stems from the operating conditions of receiver tubes, including low-power amplifiers when used in the radar band. As already indicated in Sections 10-4 and 10-5, it is expressed primarily by a reduction in gain per stage. While, in the meter and decimeter wavelengths, amplification in accordance with the frequency of received signal presents no serious difficulties, the centimeter band requires the use of traveling-wave tubes (see Section 10-9) or else the idea of high-frequency amplification must be discarded completely in view of the practical impossibility of satisfactorily matching the plate load of the tube to its resistance. The various assemblies of a receiver, working on the frequency of the received signal, do not always employ oscillatory circuits with lumped parameters. Instead it becomes necessary to resort to segments of long lines and coaxial-cylindrical cavities or cavity resonators.

It is also necessary to take into account the effects of interelectrode capacitances and inductances of tube leads. As regards atmospheric and industrial noises, their level is considerably reduced in the radar band due to the fact that the amplitudes of noise harmonics decrease continuously as their number increases, becoming quite small in the band in which signals receivable by radar are located. This is one of the most important factors requiring that the set noises of radar

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sets be reduced as far as possible, thus limiting its ultimate sensitivity.

Section 10-7. Block Diagram of the Radar Receiver

Radar receivers may display a great variety of circuit designs, including straight amplification networks. However, the majority of radar sets are of the

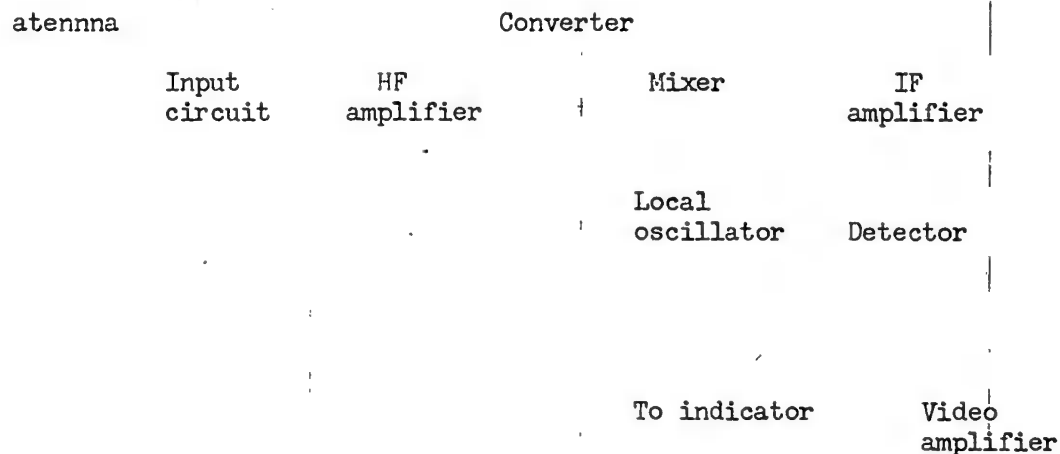


Fig.10-7 - Basic Block Diagram of a Superheterodyne
Radar Receiver

superheterodyne type.

Figure 10-7 presents a simplified typical block diagram of a superheterodyne radar receiver. It shows only the most essential assemblies:

- 1) The input circuit whose basic purpose is to match the receiver input to the HF channel of the antenna;
- 2) The HF amplifier, tuned to the frequency of the received echo signal;
- 3) The frequency converter, consisting in turn of a mixer and a local oscillator and providing conversion from the frequency of the incoming signal to an intermediate frequency more suitable for amplification;
- 4) The IF amplifier;
- 5) The detector converting IF radio pulses into video pulses;
- 6) The video-frequency amplifier, amplifying the video pulses to the level

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required for becoming visible on the indicator.

On wavelengths longer than 30 cm, radar receivers usually contain one or two HF amplifier stages, five to seven IF amplifier stages, and two to three video-amplifier stages. The reason so many stages are required lies in the difficulty of obtaining high gain, due primarily to the use of very short waves and secondly to the broad band over which amplification has to be provided.

The diagram shown in Fig.10-7 is generalized and therefore cannot correspond to the actual electric circuits of all types of radar receivers. Thus, in the centimeter band, for example, the receiver input often begins directly at the converter (the UHF stages are lacking). On the other hand, it is not uncommon for the design of a radar receiver to include auxiliary components lacking in the generalized block diagram of Fig.10-7. Such components include systems for automatic frequency trim, automatic volume-control circuits, noise-suppression systems, etc.

Section 10-8. Input Circuits

The main problem with radar receiver input circuits is that of providing maximum signal-to-noise ratio at the grid of the first tube in the receiver. The solution of this problem consists in obtaining the maximum input signal at the grid of the first tube in the receiver and in matching the resistances. Therefore, the input circuit is to be regarded as a resistance transformer matching the input impedance of the HF channel of the set to the input impedance of the first tube in the receiver.

In stations operating in the decimeter and shorter wave bands, the input circuit is an integral portion of the antenna transmit-receive switch, previously discussed in Section 6-11.

Figure 10-8 illustrates a typical input circuit of a 10-cm receiver. Matching of resistances here takes place at several points of primary importance:

a) Matching the waveguide of the main line of the waveguide channel to the coaxial line leading to the crystal mixer is accomplished by means of a cavity

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resonator, transforming the resistance when the arrester is not broken down;

b) Matching the local oscillator output so as to enable it to function on a load resistance of optimum magnitude, is accomplished by means of a tuning plunger;

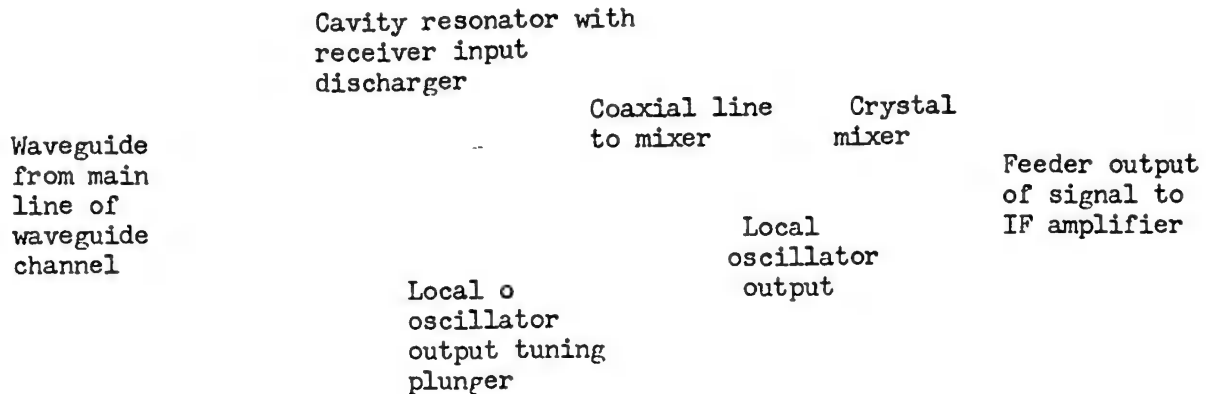


Fig.10-8 - Typical Input Circuit for 10-Centimeter Receiver

c) Matching the input resistance of the first tube of the IF frequency transformer with the line characteristic of the output feeder from the crystal mixer.

A reflex klystron serves as local oscillator in the centimeter wavelengths (see Section 10-15), while a low-power lighthouse tube, combining in its design the tube proper and coaxial-cylindrical cavities in the form of plate and cathode circuits, is used in the decimeter band (Lighthouse tubes are described in Section 11-9). Coaxial-cylindrical cavities are also used as antenna channel matching transformers.

The line characteristic of the feeder from the crystal mixer is matched to the input impedance of the first tube of the amplifier by means of a quarter-wave coaxial line (Fig.10-9) with predetermined line characteristic, constituting a feeder transformer (for intermediate frequency). The line characteristic of this transformer has to be $\rho_t = \sqrt{\rho_f R_{inl}}$, where R_{inl} is the input impedance of the first tube for intermediate frequency.

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For one-meter and longer waves, amplification at the frequency of the received signal is now universal practice. The local oscillator and the frequency converter (mixer) are taken out of the antenna input channel and become an integral part of the receiver set, which thus starts with the HF amplifier (Fig.10-10). A Class II oscillatory circuit acting as a transformer is inserted at the input of the first UHF tube, and is tuned to the frequency of the received signal and connected to the antenna feeder so that the equivalent resistance of the circuit between the connection point of the feeder and the ground is equal to the feeder line characteristic.

The conditions for matching are:

$$R_e = \frac{(p_L \rho)^2}{r} = \rho_f$$

where $p_L = \frac{L_1}{L}$ is the circuit coupling factor;

ρ is the line characteristic of the circuit;

ρ_f is the line characteristic of the feeder; and

r is the loss resistance of the circuit

Feeder from crystal mixer

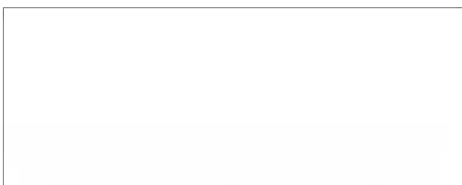
First Tube of
IF amplifier

Quarter-wave coaxial
matching line

Fig.10-9 - Matching to Input of the First Tube of IF Amplifier in Receivers for Decimeter Band

Section 10-9. High-Frequency Amplifiers

We know that the gain of a stage is:



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$$K = \frac{U_{m \text{ out}}}{U_{m \text{ in}}} = \frac{I_{ma} R_e}{U_{mg}}$$

In turn, the amplitude of the plate current of an amplifier tube is

$$I_{ma} = S_d U_{mg}$$

in which s_d is the dynamic transconductance under the operating conditions of amplification:

R_e is the equivalent resistance of the plate load;

U_{mg} is the amplitude of swing.

First cascade of HF
amplifier

R_{in}

From antenna

Fig.10-10 - Typical Input Circuit of Receiver for
Very Short Waves

In the radar band, the resistance to the plate load is so strongly shunted by the input resistance of the next stage that R_e may be taken to approximate R_{in} . The equation for the gain of the HF amplifier stage will then be

$$K = S R_{in}$$

From this we arrive at the conclusion that an increase in the gain of the HF amplifier stage requires tubes with high transconductance and high input impedance.

Bantam pentodes are used for HF amplification on waves of the order of several
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meters in length.

A typical HF amplification circuit, using a 6Zh1P circuit, is shown in Fig.10-11.

This circuit employs parallel feed of the tube plates, a resistance R_{ba} being used in place of an HF blocking choke. The tube-filament feed circuits are blocked by capacitances and small chokes, blocking the path of HF currents. Sometimes chokes are omitted from the filament circuit, being replaced by a coiling of the filament wires into a tight spiral.

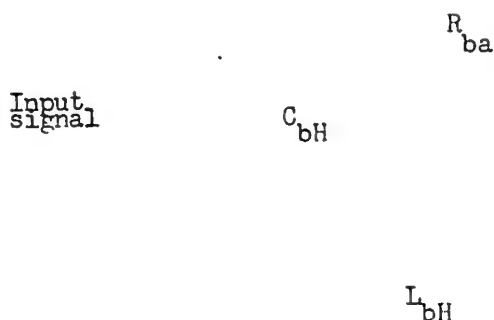


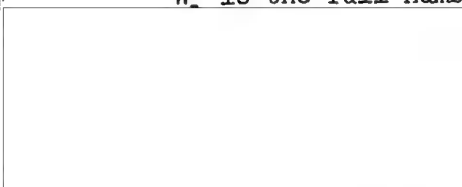
Fig.10-11 - HF Amplifier Using a 6Zh1P Pentode

In order to reduce the shunting effect of the input impedance of the next stage on the plate circuit of the preceding stage, the circuit is only partially connected to the grid circuit of the next tube. This causes the circuit to become a matching resistance transformer. The effective load on the amplifier tube plate circuit, and consequently the gain of a stage wired in this manner, depend on the coupling factor of the circuit:

$$m = \frac{W_2}{W_1} = \frac{u_2}{u_1}$$

where W_2 is the number of turns of the circuit coil coupled to the grid circuit of the next tube;

W_1 is the full number of turns in the circuit coil;



u_2 and u_1 are, respectively, the voltage delivered to the grid of the following tube, and the total voltage in the circuit.

At very low values of m , the shunting effect of the input impedance is negligible, and a comparatively high voltage is developed in the circuit, of which a very negligible fraction is utilized so that the gain of the stage is low. At m values close to unity, the circuit voltage is delivered almost entirely to the grid circuit of the following tube, but its input impedance shunts the circuit excessively so that the gain of the stage is again low.

Obviously there must be some optimum value at which the gain is at a maximum. It can be demonstrated that

$$m_{\text{opt}} = \left(\frac{W_2}{W_1} \right)_{\text{opt}} = \sqrt{\frac{R_{\text{in}}}{R_e}}$$

where R_{in} is the input resistance of the next stage;

R_e is the equivalent resistance of the plate circuit network.

This yields the following as maximum amplification factor:

$$K_{\text{max}} = \frac{1}{2} S \sqrt{R_e R_{\text{in}}} \quad (10-28)$$

In the very short wave band it is difficult to obtain an R_e of more than 5×10^3 ohms. ~~Therefore, if the input impedance of the amplifier pentode,~~ operating at the frequency of the signal, is about $R_{\text{in}} = 350$ ohms and the transconductance S is 5 ma/v, the maximum gain of the stage will be, in accordance with eq.(10-2),

$$K_{\text{max}} = \frac{5 \times 10^{-3}}{2} \sqrt{5 \times 10^{-3} \times 350} \approx 3.3$$

Thus, the gain per cascade is quite small, even in the very short wave band.

Shortening of the wave makes it necessary to dispense with pentode amplification, because of the increasing noise level, and to use triodes instead. However, the usual triode coupling employing a grounded-cathode circuit, cannot be used in

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view of the high possibility of self-oscillation due to back-coupling of the plate circuit with that of the grid across the plate-grid capacitance and the inductance of the cathode input.

Back-coupling between the plate and grid circuits in the 1-1.5 m wavelengths is reduced by means of grounded-grid triode HF amplifiers (Fig.10-12), invented by M.A.Bonch-Bruyevich. The amplifying voltage is supplied to a portion of the cathode circuit and is connected between grid and ground, as usual, while the output voltage is taken off a circuit cut in between the plate and the grounded grid. Thus, the grid and plate circuits are separated by a shield in the form of the grounded grid, eliminating feed-back between these circuits and greatly weakening the possibility of self-excitation.

In constructing the equivalent circuit of an amplifier employing a triode and grounded grid, it must be borne in mind that the voltage applied to the grid is effective twice: as amplified by a factor of μ and as directly cut into the

u_{in}

Fig.10-12 - Triode Amplifier with Grounded Grid

circuit of the series-coupled plate and grid circuits (Fig.10-12b).

In accordance with the equivalent network, the plate current is

$$\bar{I}_a = \frac{(\mu + 1) \bar{U}_g}{R_i + R_a}$$

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and the gain per stage is

$$K = \frac{\bar{U}_{out}}{\bar{U}_{in}} = \frac{(\mu + 1) R_a}{R_i + R_a} \approx \frac{\mu R_a}{R_i \left(1 + \frac{R_a}{R_i}\right)}$$

In view of the fact that

$$\mu \gg 1, \text{ while } R_i \gg R_a$$

we obtain:

$$K \approx S R_a \quad (10-29)$$

The input impedance of a stage with grounded grid

$$R_{in} = \frac{\bar{U}_{in}}{\bar{I}_{in}} = \frac{\bar{U}_g}{\bar{I}_a} = \frac{R_i + R_a}{\mu + 1} \approx \frac{R_i + R_a}{\mu}$$

When $R_i \gg R_a$, we obtain

$$R_{in} = \frac{1}{S} \quad (10-30)$$

To increase the gain, incomplete coupling of the circuit either on the input or the output side is used. Here again there is an optimum coupling factor:

$$m_{opt} = \sqrt{\frac{R_{in}}{R_a}}$$

which, in accordance with eq.(10-28), provides the maximum amplification factor:

$$K_{max} = \frac{1}{2} S \sqrt{R_{in} R_a} \quad (10-31)$$

The theory of amplification by grounded-grid triodes has been worked out in detail by V.I.Siforov, and A.A.Kolosov.

Because of their very wide frequency range, the circuits of HF amplifiers for radar receivers, as distinct from broadcast-band receivers, have fixed tuning, and the transmitter is not trimmed when frequency drift occurs. Moreover, sets operating close to the limit of the very short wave band do not even contain capacitors

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in the form of coils. The capacitances of the circuits are provided by the stray capacitances of the network.

To amplify oscillations when the frequency of the received signal is of the order of 3000 mc and more (microwave band), a special category of vacuum tubes is used: the traveling-wave (TW) tube. The diagram of this type of tube is shown in Fig.10-13.

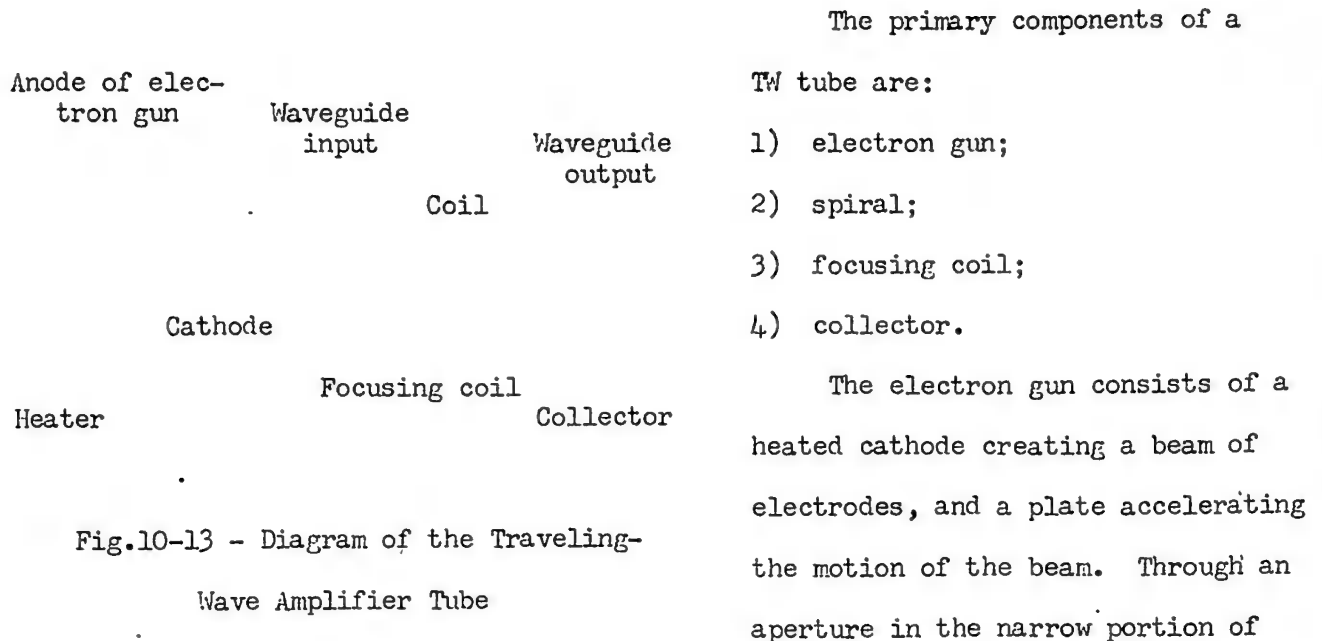


Fig.10-13 - Diagram of the Traveling-Wave Amplifier Tube

the anode, the beam travels by inertia along a closely coiled wire spiral and is picked up by the collector, with the same potential as the anode (+1000 to 1500v).

A long focusing coil occupying the neck of the tube serves to compress the beam in the direction of the axis of the spiral.

The signal being amplified enters through the input waveguide and induces a traveling electromagnetic wave, propagating along the coils of the spiral at the velocity of light. The longitudinal component of this velocity, which is the speed at which the wave moves along the axis of the spiral, is smaller than the velocity of light by the same factor by which the spacing of the windings of the spiral are smaller than the length of one of its turns.

The traveling electromagnetic wave acts on the electron beam. If the velocity

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of the electron beam is somewhat greater than the longitudinal velocity of the electromagnetic wave, the beam is retarded and yields energy to the electromagnetic wave. As a result, the voltage of the field induced in the outgoing waveguide is 10 to 15 times as great as in the incoming.

Amplification with a TW tube ensures a relatively low level of set noise and, in addition, does not set up a tuned amplifier. A TW tube, like a waveguide, functions over a very wide band and does not require tuning to the frequency of the incoming signal.

Section 10-10. Frequency Converters in Radar Receivers

The problem of frequency conversion in the radar receiver lies in transforming a radio pulse at the frequency of the incoming echo signal to a pulse of a lower, intermediate, frequency, useful for further amplification relative to the noise level and for obtaining a band wide enough to provide a permissible distortion of the shape of the echo signal.

As in all superheterodynes, the radar receiver includes a mixer actuated by the frequency voltage of the incoming echo signal f_s , and the frequency voltage f_h of the heterodyne, a low-power local oscillator with direct drive, also constituting a standard component of all superheterodyne receivers.

According to the classic theory of frequency conversion, developed by Prof. V. I. Siforov, the effect of two different frequencies, f_s and f_h , on a voltage mixer causes significant changes in its parameters, of such a nature that the transconductance of the mixer acquires a significant component S_{if} , representing an intermediate frequency $f_{if} = f_s - f_h$. If the mixer is loaded with a load tuned to the frequency f_{if} , this load can produce an IF voltage across a resistance R_a .

Like all stages, the mixer has a given gain factor in intermediate frequency:

$$K_{if} = S_{if} R_a \quad (10-32)$$

As the conversion transconductance is only a portion of the total transconductance S of the mixer characteristic, the gain of a tube when used as a mixer is always less than that of the same tube when used as an amplifier.

Both the incoming signal frequency and the local oscillator frequency exceed the frequencies of ordinary radio broadcast receivers. Therefore, the noise levels and input impedances of ordinary converter and mixer tubes (pentagrids, hexodes, etc.) make them entirely unsuited to use in radar receivers.

In very short wave receivers, the same types of tubes used for amplification are used for conversion, i.e., special pentodes or triodes. However, while in broadcast-band receivers, the incoming signal voltage and the local oscillator voltage are received at different tube grids to reduce the mutual effects of signal and local oscillator circuits, in radar receivers the object of noise reduction and increase of amplification is served by connecting the converter, the signal voltage and the local oscillator voltage to the same grid of the mixer tube or, in other words, single-grid conversion is employed.

Figure 10-14 illustrates one diagram for a single-grid mixer for waves at the end of the very short wave band. It also shows a local oscillator assembled around a separate tube L_h in the Hartley arrangement. The input circuit of the mixer tube is tuned to the frequency of the signal and is, as a consequence, considerably mistuned relative to the local oscillator. The need for an input circuit capacitance is met by the input capacitance of the mixer tube, the capacitance of the wiring and the capacitance C_L of the circuit coil L . In terms of design, the circuit is executed as a coil with a brass or carbonyl trimming core. Incomplete coupling of the input circuit permits matching to the plate load of the previous stage, the HF amplifier. The tuning of the mixer input circuit is fixed as is that of all the HF circuits. When frequency drift of the incoming signal occurs, only the local oscillator circuit is tuned, by means of its capacitor C_H , to provide constant intermediate frequency.

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The automatic grid bias of the mixer created by the cathode current across the resistance R_{mx} , is set close to the mixer-tube grid cutoff voltage so as to operate

to HF amplifier

to IF amplifier

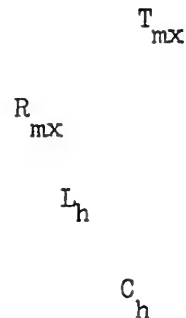


Fig.10-14 - Mixer and Local Oscillator of Very Short Wave Receiver

~~in the range of maximum variation in total mixer-tube transconductance.~~

Local oscillators for the microwave band are usually assembled around light-house tubes, while the mixer tube is a diode which, while providing no amplification has considerably more input impedance and creates less noise than a triode or pentode.

Figure 10-15 shows a typical diode mixer diagram. The signal and local oscillator voltages are impressed in series onto the mixer-tube circuit, as with pentode or triode mixers. The IF voltage is taken from a circuit wired as a coil L_{if} . To provide the most favorable operating conditions for the mixer diode and to obtain maximum conversion effect, a supply source E_{mx} is cut in.

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At wavelengths of less than 50 cm (particularly in the microwave band), crystal detectors are used as mixers. The coupling diagram of a crystal mixer has already

Signal f_s

To IF amplifier

Local oscillator
voltage f_h

E_{mx}

L_{iF}

f_{if}

Fig.10-15 - Typical Diagram of Diode Mixer

been illustrated in Fig.10-8. The operation of the crystal detector is based on the fact that the contact between the conductor and the semiconductor crystal

a)

b)

i_{cr}

Germanium

Silicon

Fig.10-16 - A Crystal Detector

a - Design; b - Performance curve

conducts in one direction only. Silicon or germanium crystals are used.

The electric characteristic of a crystal detector is sufficiently linear (Fig.10-16b). The slope of the curve in the region of positive voltages determines the direct resistance of the contact, while the slope in the region of negative

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voltages governs the back resistance of the contact. The normal direct resistance of silicon crystals is in the 50 to 500-ohm range, and the back resistance in the 1000 to 20,000-ohm range. In view of the presence of a design capacitance of the order of $1 \mu\text{f}$, shunting the contact layer, the input impedance of the crystals is complex at high frequencies and has a capacitive reactance. Its active component is of the order of 20 to 60 ohms (usually, 50 ohms). The output impedance of the crystal at intermediate frequencies is also complex (capacitive reactance), the active component being of the order of 200 to 600 ohms (usually 300 ohms).

In design (Fig.10-16a), the crystal detector is built into a ceramic body with brass terminals (5) and (6), the former being connected to the inner wire of the outgoing coaxial line to the IF amplifier and the second to the inner wire of a coaxial line, simultaneously affected by the signal and heterodyne frequency fields (Fig.10-8).

The crystal (1) is soldered to the adjusting screw (4). A tungsten catwhisker (2), fastened to the lead terminal (6), serves as contact conductor. The closeness of contact between crystal and wire is controlled by the screw (4). To prevent the parts from moving, the internal space in the ceramic body (3) is filled with wax.

The advantages of the crystal mixer relative to the diode are as follows:

- 1) Considerably lower input capacitance, significantly simplifying the design of both the receiver input circuit and that of the intermediate-frequency amplifier;
- 2) The linearity of the curve at low voltages, ensuring effective operation when the input signal is weak;
- 3) The reduced noise level, due to absence of an emitter cathode.

The main drawback of the crystal mixer is its sensitivity to overloads, due to the small contact area between the wire and the crystal.

Section 10-11. Determination of Intermediate Frequency

In determining the intermediate frequency of a radar receiver one must accept a compromise solution that will satisfy, to the best possible degree, a number of contradictory requirements.

1. The intermediate frequency should not be in the band of powerful ultrashort-wave broadcasting stations, to prevent significant interference with reception of echo signals.

2. The intermediate frequency must be low enough to permit weakening of the noise level of the IF amplifier and to ensure minimal detuning of the circuits due to parameter spread when tubes are changed since, at the higher frequencies, the interelectrode capacitances of the tubes comprise a greater portion of the total circuit capacitance than is the case at lower frequencies.

3. The intermediate frequency must be sufficiently low to ensure a high gain per stage, permitting a reduction in the total number of IF amplifier stages.

4. The intermediate frequency must be sufficiently high to weaken interference by stations in the image channel. The image channel operates at a frequency that, as distinct from the frequency of the echo signal, is not higher but lower than the local-oscillator frequency by the value of the intermediate frequency. Therefore,

on passing through the mixer, the mirror station also yields an intermediate frequency and is consequently magnified in the intermediate frequency amplifier. The higher the intermediate frequency, the greater will be the difference of the frequency of the incoming signal from that of the mirror station, and the greater will be the degree to which the input circuit of the receiver is out of tune from the station in the mirror channel.

5. The intermediate frequency must be sufficiently high to prevent its voltage from coinciding with the pass band of the video amplifier and thus be amplified. This condition is satisfied if the frequency is five times greater than that of the video amplifier pass band, i.e., it must meet the condition:

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$$f_{if} \geq 5\Delta F_{\text{video freq}}$$

where $\Delta F_{\text{video freq}}$ is the pass band of the video amplifier.

6. The upper limit of the intermediate frequency is determined by the resistance of the stage to direct drive due to feedback across the plate control-grid capacitance. As demonstrated by Prof. V.I. Siforov, the stable gain factor of a stage is

$$K_{\text{stable}} \leq 0.42 \sqrt{\frac{S}{\omega_{if} C_{ag}}}$$

An increase in the intermediate frequency ω_{if} is accompanied by a decline in the factor K_{stable} and requires an increase in the number of stages in the IF amplifier.

The frequencies from 15 to 100 mc represent a compromise solution of the problem. The intermediate frequencies most frequently used in radar receivers are three: 15, 30, and 60 mc.

Section 10-12. Intermediate-Frequency Amplifiers

The most commonly used intermediate frequencies in radar receivers are 30 or 60 mc. Although these frequencies already extend into the ultrashort-wave band (5 and 10 m), electron transit time does not yet make itself felt and there is no significant drop in tube input impedance. The special difficulties involved in operating on the ultrashort-wave band are not encountered in amplifying intermediate frequencies.

Most of the total amplification in the receiver, which is of the order of several millionfold, occurs in the intermediate-frequency amplifier. It is important to note in this connection that IF amplifier circuits are determined for the most part not by the operating wave of the receiver but by the specifications for frequency range, i.e., for quality of reproduction of the pulse envelope, and by

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the total gain factor in the intermediate frequency. It should be added that, if amplification in the intermediate frequency precedes HF amplification, the IF amplifier stages need meet no special requirements as far as noise is concerned, since set noise under these conditions is determined by the HF amplifier stages, the mixer and the local oscillator. In the case where the IF amplifier is the first amplifier stage of the receiver, the noise it creates requires very serious attention, a fact that is naturally reflected in the IF initial stage amplifier diagrams.

Thus, the decisive factor in determining the IF amplifier diagram is the IF pass band required.

As we know from radio engineering, the equivalent resistance of a class I parallel circuit is:

$$R_e = \frac{\rho^2}{r} = \frac{\rho}{r} \times \rho$$

or

$$R_e = \frac{\rho}{d} \quad (10-33)$$

where $\rho = \sqrt{\frac{L}{C}}$ is the characteristic impedance of the circuit;

$d = \frac{r}{\rho}$ is the attenuation of the circuit.

We also know that the pass band of the circuit is

$$\Delta F = df_o$$

where f_o is the resonant frequency of the circuit. From this we derive

$$d = \frac{\Delta F}{f_o} \quad (10-34)$$

Substituting eq.(10-34) in eq.(10-33), we obtain:

$$R_e = \frac{f_o \rho}{\Delta F}$$

Finally, considering that $\rho = \frac{1}{\omega_o C}$ and $\omega_o = 2\pi f_o$, we derive the equation for STAT

the equivalent resistance the circuit requires in order to provide the required frequency range:

$$R_e = \frac{1}{2\pi\Delta F C} \quad (10-35)$$

where C is the capacitance of the circuit.

From this it follows that, in order to obtain the greatest possible R_e (i.e., maximum gain per stage) at a given ΔF , it is necessary to reduce the capacitance of the circuit, whose minimum value is limited by the stray capacitance at the stage output, of the order of 15-25 μf . For example, to provide a band $\Delta F = 6 \text{ mc}$ the circuit required must have an equivalent resistance of

$$R_e = \frac{1}{2\pi \times 6 \times 10^6 \times 25 \times 10^{-12}} \approx 1000 \text{ ohms}$$

In order to obtain such low R_e values, the IF circuits have to be designed as coils, usually with an inductance of several microhenries, shuntable by a resistance of the order of 2-4 ohms, in order to pass the required band of frequencies.

To obtain satisfactory gain per stage at a low resistance R_e to plate load, it is necessary to use amplifier pentodes of high transconductance (7-12 ma/v) and employ a large number of intermediate frequency amplifier stages (from 4 to 9).

Amplifier networks for the intermediate frequency may be classified in two main groups: single-circuit and twin-circuit.

In single-circuit networks (Fig.10-17a), the circuit coil I_c is coupled to the grid circuit of the following stage and shunted by the resistance R_{sh} , coupled to the plate circuit of the preceding stage. Should the total gain in the intermediate frequency be comparatively small, the circuits of all IF amplifier stages are tuned to the middle of the IF pass band. However, when high IF amplification is required over a relatively wide band, single-circuit systems with symmetrically detuned pairs of stages, one tuned to a frequency below the midfrequency and the other to a frequency above this. Finally, when the band used is very wide, exceed-

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ing 15 mc, networks employing several cascades of three stages are used. The first stage in the cascade of three is tuned to a frequency below the midfrequency, the

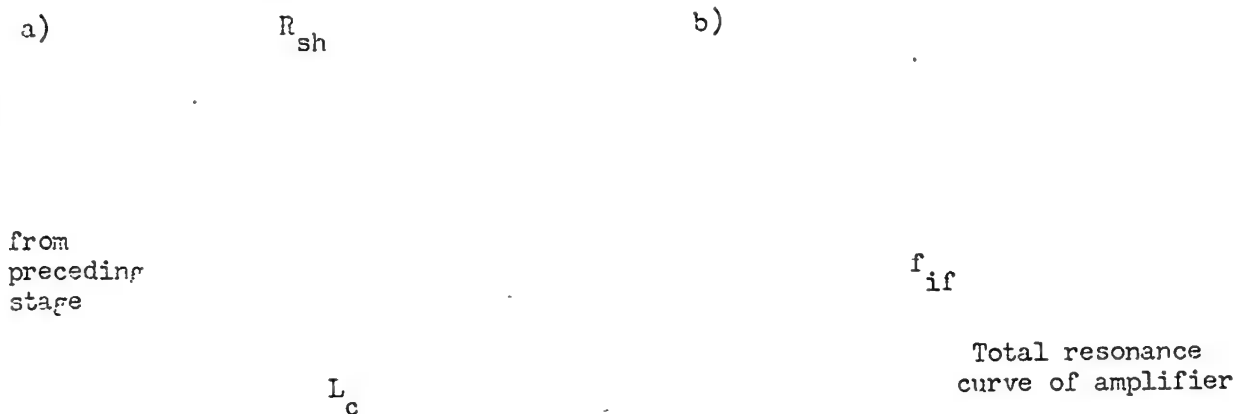


Fig.10-17 - Single-Circuit IF Amplifier

a - Diagram; b - Resonance characteristic of
a system of three detuned stages

second only to the midfrequency itself, and the third to a frequency somewhat above it.

Figure 10-17b depicts the resonance curves of three cascades staggered in the manner indicated. The total resonance curve of the three cascades is found by multiplication of the ordinates of the resonance characteristics of these stages, since in a multistage amplifier the total gain is the product of the gains of the individual stages. The total specified band of IF amplification is obtained by choice of the parameters of the circuits of individual stages and of the degree of stagger. In this manner, individual stage pass bands are obtained that are sufficiently narrow to provide good amplification per stage.

If there has been no pre-amplification ahead of the IF amplifier (as in the microwave receivers of gun-ranging sets), the networks of the first stages of the IF amplifier are affected somewhat. In view of the fact that it is necessary to reduce the noise level to a minimum, the first two stages use triodes instead of pentodes. The first triode is coupled on a grounded cathode network, since this

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provides for less noise. In order to reduce or eliminate self-excitation due to grid-plate capacitance, the plate load of the first stage is intentionally reduced at the expense of its gain. This is done by connecting a second triode in the grounded grid coupling. It should be remembered that the input impedance of a cascade with grounded grid is virtually independent of the frequency. If the impedance is fairly large at 200-300 mc, then at 30 to 60 mc, on the other hand, it becomes small relative to a network with grounded cathode. The result of the strong shunting action of the input impedance of the second stage causes the first-stage gain to decline, so that self-excitation becomes more difficult.

Figure 10-18 illustrates such a grounded cathode - grounded grid network. From the mixer output, the signal is impressed on the first tube grid across the input transformers L_s and L_{amp} , matching the output impedance of the mixer to the input impedance of the IF amplifier. The diagram also shows a device used for

from mixer

to next IF amplifier stage

L_s

L_{amp}

Ch_1

Ch_2

crystal current

Fig.10-18 - Grounded Cathode - Grounded Grid Network

controlling the direct component of the crystal-mixer current. The filters C_1 , Ch_1 and C_2 protect the device from the effects of intermediate frequency. The choke Ch_2 protects the HF component from contact to ground.

Two-circuit diagrams of IF amplification (Fig.10-19) include, in the plate circuit of each stage, a band filter of two inductively-coupled circuits, shunted

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by the resistances $R_{sh a}$ and $R_{sh g}$, respectively, to provide the necessary pass band without significantly reducing the gain per stage. The coupling between the circuits is usually set at the critical, when the resonance characteristic still has only one maximum. Less frequently, this characteristic is set at above the critical level in order to obtain a two-peak filter resonance characteristic.

Where two-circuit diagrams are used, pairs of IF amplifier stages, symmetrically staggered relative to the middle of the pass band, are used.

Section 10-13. Detection

The purpose of the detector in a radar receiver is to convert the IF radio pulses into the video pulses required, after further amplification, for direct

reading of the target coordinates on indicator screens.

In general, detectors have to satisfy the following conditions in order to meet this requirement:

from preceding
stage of IF
amplifier

- a) Undistorted detection of IF signals, i.e., undistorted reproduction of the envelope of the amplified and converted radio pulse reflected from the target;
- b) Provision of maximum transfer

Fig.10-19 - Twin-Circuit IF Amplifier

constant for the amplitude of the detected pulse;

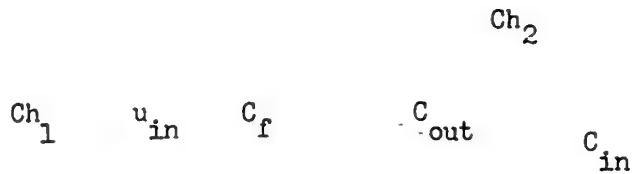
- c) Provision of minimum penetration of the IF voltage into the video amplifier grid of the first stage;
- d) No reduction in the signal-to-noise ratio established at its input by the preceding stages of the receiver.

In principle, all types of detection known to radio engineering may be employed in radar. However, diode detectors are preferred, since they provide the broadest STAT

dynamic range or, in other words, the widest band of detectable radio pulse amplitudes resulting from differences in the target range.

a)

video amplifier first stage



b)

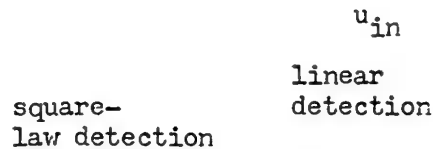


Fig-10-20 - Diode Detection

a - Typical network; b - Detection characteristic

Figure 10-20b illustrates the detector characteristic of a diode detector.

Depending on the swing of the detectable signal, passing from the IF amplifier output to the detector input, detection is classified as square-law, which embraces the lower curvilinear (approximately square-law) portion of the detector curve, and as linear detection, embracing the upper (linear) portion of the detector characteristic.

In square-law detection, the amplitude of the pulse at the detector output is approximately proportional to the square of the amplitude of the IF pulse at the detector input. Therefore, it is desirable that detection of weak signals be in the square-law band (and the IF amplifier is designed accordingly) in order to arrive at the signal-to-noise ratio that is so important in this connection.

In linear detection, the pulse amplitude at the detector output is approximately proportional to the first power of the IF pulse amplitude at the detector input. Therefore, when hookups providing for noise suppression are used, it is better that detection occur in the linear interval.

Figure 10-20a presents a standard diode detector diagram.

The choke Ch_1 is designed to prevent short-circuiting of the IF current of the IF amplifier final stage. The C_f filter and Ch_2 are primarily designed to ground the IF component at the detector output and secondly to block the path of IF currents across the load resistance R_n and beyond that to the video channel of the receiver. The filter choke Ch_2 has a natural frequency equal to the intermediate frequency and behaves like an antiresonant parallel circuit connected in series or, in other words, like an IF wavetrap. Thanks to this entire system of IF filters, only the video frequency component of the detected radio pulse spectrum or, in other words, the video pulse voltage is discriminated at the detector load resistance R_n .

In the majority of cases, negatively polarized video pulses are discriminated by the load resistance. This permits limitation of the amplitude of the echo signals in the first stage video amplifier, with the object of eliminating the possibility of overload of the succeeding video amplifier stages.

The analysis of the detector parameters consists in determining the load resistance. This is based on:

- a) The equivalent load capacitance

$$C_e = C_f + C_{out} + C_{in}$$

where C_f is the capacitance of the IF output filter;

C_{out} is the output capacitance of the detector diode;

C_{in} is the input capacitance of the first stage of the video transformer;

- b) the required video channel pass band $\Delta F_{vid f}$.

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The detector load resistance, like the equivalent resistance of the IF amplifier circuit, is determined from the equation:

$$R_n \leq \frac{1}{2\pi\Delta F_{vid} f C_e} \quad (10-36)$$

The transfer function of diode detectors in radar receivers is in the range of 0.1 to 0.7.

Section 10-14. Video Amplifiers

The purpose of video amplifiers in radar receivers is to amplify detected signals to a level adequate for normal indicator operation. It should be mentioned that this level is not always determined by the operating voltage required for normal tube operation. Very often the indicator contains its own supplementary video amplifier. In this case, the receiver has to amplify the video pulse to a level adequate for normal operation of the auxiliary video amplifier of the indicator (approximately to 5 or 8 v). Sometimes the indicator is as much as some tens of meters distant from the receiver, and the outgoing video signal is transmitted by coaxial cable. In this case, the receiver video amplifier has to work on a low-resistance matched load $R_n = \rho_f$, where ρ_f is the line characteristic of the connecting feeder.

In the vast majority of cases, video amplifiers are designed for amplification across resistances. Figure 10-21a presents a generalized diagram of such an amplifier. As we know, the frequency response curve of such an amplifier (Fig. 10-21b) has two steep slopes located

a) in the low-frequency band of the amplified signal, due to the very high resistance of the separating capacitor C_g , at low frequencies;

b) in the upper frequencies, due to the output capacitance of the stage and the dynamic input capacitance of the following stage which, at high frequencies, is forcefully shunted by the resistance R_a of the video-amplifier plate load.

The pass band of resistance amplifier does not, in accordance with the diagram STAT

in Fig.10-21a, exceed

$$\Delta F_{vid f} = \frac{1}{2\pi R_n C_e} \quad (10-37)$$

where $C_e = C_{out 1} + C_{wiring} + C_{in 2}$ denotes the equivalent capacitance of the diagram.

a)

$C_{out 1} + C_{wiring}$

u_{out}

$C_{in 2}$

b)

steep slope in low
frequencies

steep slope in high
frequencies

Fig.10-21 - Resistance Amplifier

a - Generalized circuit; b - Frequency response

One of the major requirements to be met by a video amplifier is that the build-up and decay time of the outgoing video pulse be brief. The briefer that period, the higher will be the accuracy of ranging. If no special steps are taken to meet this requirement, the presence of even minimal stray capacitances make it necessary in the conventional resistance amplifier design, to reduce the resistance to plate load, i.e., to widen the pass band in accordance with eq.(10-37); this results on a reduction in the video amplifier gain per stage. However, the requirement that the gain be increased necessitates, on the other hand, an increase in plate-load resistance, and this, in accordance with eq.(10-37), leads to a reduction in the pass band and is accompanied by an increase

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in the rise and decay time of the output pulse.

In order for an adequately uniform frequency response to be obtained throughout the video amplifier pass band, determined as previously described (see Section 10-6) or, in other words, in order to assure low build-up and decay time in the face of high plate load resistance, the video amplifier diagram should include components for correcting the frequency response in the high as well as in the low frequencies.

Since the lowest frequency in the spectrum of the repeated video pulse is determined by the repetition rate, and the steep response slope at the low frequencies usually becomes apparent only in the interval below 200 cycles, it is not

a)

L_{comp}

b)

with compensation

without compensation

relative
gain

Fig.10-22 - Video Amplifier with HF Compensation

a - Diagram; b - Frequency response

necessary to provide low-frequency correction when the pulse repetition rate exceeds 200 cycles.

Thus, when the frequencies are sufficiently high, frequency response correction need be repeated only in the high frequencies. A video amplifier meeting this

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requirement would have the diagram shown in Fig.10-22a, and the frequency response shown in Fig.10-22b. The simplest type of compensating circuit is used here, in the form of an inductance L_{com} coupled in series to a resistance R_a of the amplifier plate load. The compensating inductance and the equivalent capacitance $C_e = C_{out 1} + C_{wiring} + C_{in 2}$ form an oscillating circuit of high attenuation, providing a smooth rise in frequency response at the resonant frequency of this circuit.

The amount of inductance is so selected as to yield resonance at a higher frequency than $f_{e \text{ slope}}$, at which a steep slope below the 0.7 level begins (Fig.10-22b). This compensation widens the pass band relative to the cutoff frequency $f_{e \text{ slope}}$ by a factor of $\sqrt{2}$, where the inductance, subject to correction, can be calculated from the equation

$$L_{comp} = 0.25 R_a C_e$$

The diagram of a video amplifier with compensation only at low frequencies (to simplify the drawing) is shown in Fig.10-23a. The compensating low-frequency circuit consists of a resistance, R_{comp} and a capacitance C_{comp} , cut into the plate circuit. In the low-frequency range, the resistance of the capacitor C_{comp} is of adequate magnitude and does not shunt the resistance R_{comp} .

Consequently, at low frequencies it may be considered that a supplementary resistance R_{comp} is connected in series with the resistance R_a . As a result of the increase in the total resistance of the plate load of the amplifier tube, there is an increase in the gain factor at low frequencies, with the result that the steep slope in the amplifier frequency response at low frequencies is eliminated. At intermediate and high frequencies, the resistance R_{comp} is so strongly shunted by the capacitance C_{comp} that it has no significant effect on the gain per stage. Satisfactory compensation of the frequency response in the low-frequency band is usually obtained if $R_{comp} \approx 5R_a$ and, in addition, if the following condition is satisfied:

$$R_a C_{comp} = R_g C_g \quad (10-38)$$

The input resistance of the indicator circuits contains a large capacitance factor, strongly shunting the load resistance of the final stage. It is, therefore,

a) C_{comp} R_{comp}

video pulse
to be
amplified

b) with compensation

without compensation

200 cycles

Fig.10-23 - Video Amplifier with Low-Frequency Compensation

a - Diagram; b - Frequency response

desirable that the load resistance of this stage be low. In this case, the build-up and decay time of the output pulse will be small and the pass band will be of adequate width even without any compensation of frequency response. This is of particular importance when the video amplifier is working on a connecting coaxial cable some tens of meters in length, which, for purposes of matching, is loaded at the indicator side with a resistance equal to the characteristic impedance. No significant shunting of so small a load resistance $R_k = \rho$ of the indicator input capacitance will occur, even if the frequency band is quite broad.

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For effective operation against a low load resistance, the output stages of the video amplifier are usually designed in the form of the cathode follower shown in Fig.10-24a (in the case of a long cable) or in Fig.10-24b (when a short cable functions as a small shunting capacitance). In the cathode-follower network, the load is small and is coupled into the cathode circuit in any manner, i.e., between cathode and ground. The output voltage U_{out} , separated by the resistance R_k , is reversed in the grid circuit counter to the input voltage U_{in} , which signifies the presence of a negative (antiphase) couple in the cathode follower. Due to the negative feedback, reducing the resultant grid voltage, the cathode follower is unable to function as a voltage amplifier. In fact, if we neglect the grid current relative to the plate current, it may be considered that the output voltage is

$$u_{out} = i_a R_k$$

The plate current is determined by the transconductance S and by the resultant grid voltage which, in view of the presence of negative feedback, will equal

$$u_g = u_{in} - u_{out}$$

so that

$$i_a = S u_g$$

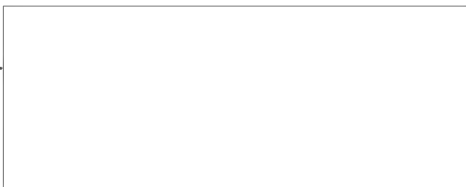
and

$$u_{out} = u_g S R_k = (u_{in} - u_{out}) S R_k$$

or

$$u_{out} (1 + S R_k) = u_{in} S R_k$$

From this we find the gain factor of the cathode follower, as



$$K = \frac{u_{out}}{u_{in}} = \frac{SR_k}{1 + SR_k} \quad (10-39)$$

Equation (10-39) shows that the gain of a cathode follower is always somewhat less than unity in voltage. Here we note that, in the circuit of the cathode follower, the output voltage differs from that in the usual type of amplification in

a)

c)

to video amplifier
indicator grid

b)

 u_{in}

to video amplifier
indicator grid

Fig. 10-24 - Cathode Follower

a - With long cable; b - With short cable; c - Equivalent network

that it is in phase with the input voltage and "follows" any change in this voltage; this is the source of the name "cathode follower".

Equation (10-39) may be converted to a clearer form if we consider that $S = \frac{\mu}{R_i}$. As a result of this substitution, we readily obtain the equation

$$K = \mu \frac{R_k}{R_k + \frac{R_i}{\mu}} = \mu \frac{R_k}{R_k + R'_i} \quad (10-40)$$

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The formula for a conventional amplifier is

$$K = \mu \frac{R_a}{R_a + R_i} \quad (10-41)$$

A comparison of eqs.(10-40) and (10-41) shows that the cathode follower can be replaced by an equivalent standard amplifier with load resistance R_k and the following total internal tube impedance:

$$R'_i = \frac{R_i}{\mu}$$

or

$$R'_i = \frac{1}{S} \quad (10-42)$$

As indicated by the equivalent circuit (Fig.10-2/c), the resistance R'_i shunts the cathode resistance R_k so that the equivalent output resistance of the cathode follower at the points becomes

$$R_{out} = \frac{R'_i R_k}{R_k + R'_i} \quad (10-42a)$$

Thus, the cathode follower presents a very low output impedance, not in any case exceeding $R'_i = \frac{1}{S}$, which may readily be matched to the line characteristic of the cable. However, the input resistance of the cathode follower may be very high if it is operated with very low grid currents. As the input and output voltages are of the same order of magnitude in cathode follower circuits, a greater input resistance signifies a relatively smaller input power, and a low output resistance a relatively high output power. Thus, the cathode follower functions as a power amplifier, simultaneously serving to match low-resistance load with high load resistance of the tube in the stage before the final stage.

Section 10-15. Automatic Frequency Trim

Should the frequency of the transformer or the local oscillator be changed,

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the latter must be aligned so that the frequency difference $f_s - f_h$ between signal and local oscillator will always equal the intermediate frequency f_{if} of the receiver (30 mc). If the intermediate frequency changes due to transmitter or heterodyne frequency drift, the receiver will be detuned in the intermediate frequency and reception of echo signals will be interfered with. Therefore, automatic local-oscillator frequency control (AFC) systems are widely used in radar receivers.

Within the band it is designed to align, the AFC system should provide very rapid trim within the intermediate frequency even when sudden transmitter skipping has occurred, and should compensate for all instabilities in the frequency both of the transmitter and of the local oscillator itself. It should not be possible for the receiver to tune to the image frequency (see Section 10-11). The image frequency is weakened adequately ahead of the mixer input by resonance amplification of the reflected signal frequency in the HF amplifier stages.

The simplest AFC block diagram is that shown in Fig.10-25. The AFC mixer, like the receiver main channel mixer, yields an IF voltage amplified by one or two stages of the IF amplifier, when the frequency voltages of the specially weakened transmitter and local oscillator signals are applied. The amplifier output voltage acts on the discriminator. The discriminator yields a voltage whose magnitude and sign depend on the magnitude and sign of IF detuning. Guided by the control network, it then acts on the local oscillator frequency in such a fashion that the IF receiver detuning is rapidly diminished. Some amplification in the intermediate frequency is necessary if the output voltage of the automatic frequency control mixer (particularly a crystal mixer), comprising 0.2 to 0.5 volt, is to be brought up to the level of a few volts, required for the discriminator diodes to function in linear detection.

The Discriminator

Discriminators are often designed in accordance with the diagram in Fig.10-26, which is no different in principle than that of an FM discriminator. The L_1C_1

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and L_2C_2 circuits are tuned strictly to the intermediate frequency f_{if} , selected for the receiver (usually, 30 mc). The connection between the circuits is quite loose, so that the resonance frequencies of the circuits are determined only by

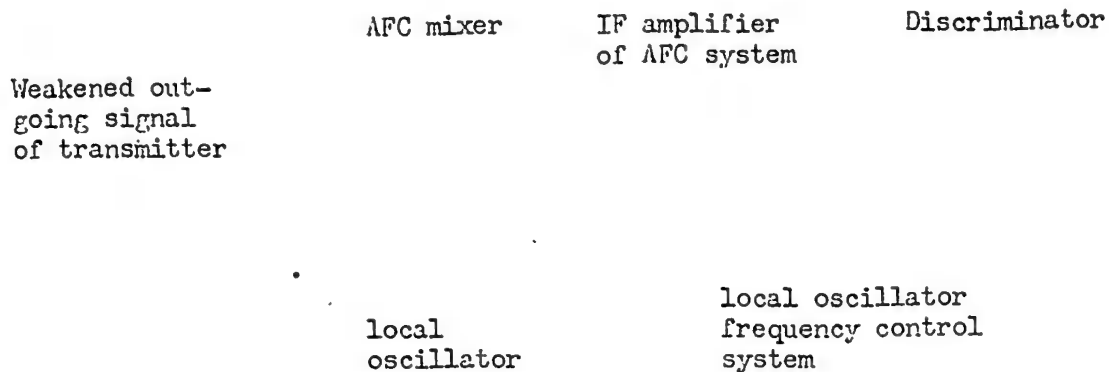


Fig.10-25 - Simplest Type of AFC Diagram

their own parameters. Both circuits have high attenuation. The resonance curves of the circuits are wide enough, and the voltage in the circuits is practically constant regardless of changes in the intermediate frequency within the automatic tuning band.

The discriminator provides automatically the amount and sign of the automatic trim voltage impressed on the control network by the local oscillator.

Inductive coupling in the coil L_2 causes induction in the first circuit of some emf \bar{E} of intermediate frequency. Regardless of the magnitude of this frequency, the voltage \bar{E}_2 at the capacitor C_2 lags behind the current \bar{I}_2 , in the second circuit by exactly 90° . Because of the presence of adequate capacitances C_3 and C_4 , the cathodes of the diodes D_1 and D_2 are at zero potential in the intermediate frequency. The lower end of the coil L_1 in the first circuit, is also of zero potential. Therefore, the voltage \bar{E}_1 on this coil is connected across the capacitor C_0 (whose resistance can be neglected) to the midpoint of the coil L_2 . Therefore, at intermediate frequency, the diode D_1 is under a voltage

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of $\bar{U}_1 = \bar{E}_1 - \frac{E_2}{2}$ while the diode D_2 is under a voltage of $\bar{U}_1 = \bar{E}_1 - \frac{E_2}{2}$.

The output voltage of the discriminator is the difference between the voltages created by the diode currents at the resistances R_3 and R_4 , as a result of the detection of IF pulses:

$$U_{out} = U_3 - U_4$$

Each intermediate frequency is of normal magnitude (30 mc), while the current I_2 coincides in phase with the emf \bar{E} and the voltage \bar{E}_1 (see the vector diagram in Fig.10-27b). The voltage \bar{E}_2 at the capacitor C_2 lags behind the voltage \bar{E}_1 by

from IF amplifier

$$U_{out} = U_3 - U_4$$

R-F choke

Fig.10-26 - Discriminator

exactly 90° . The voltages U_1 and U_2 , impressed on the diodes, are equal in magnitude. The rectified voltages \bar{U}_3 and \bar{U}_4 compensate each other completely, and the output voltage of the discriminator thus becomes $\bar{U}_{out} = 0$. Therefore, when the magnitude of f_{if} is normal, the discriminator does not affect the heterodyne frequency-control circuit.

At intermediate frequencies above normal (30 mc), the second circuit shows an inductive reactance. The current \bar{I}_2 lags behind the emf \bar{E} by a given angle φ , while the voltage \bar{E}_2 now lags behind the voltage \bar{E}_1 not by 90° , but by $90^\circ + \varphi$ (see vector diagram in Fig.10-27a). Therefore, the absolute magnitude of the voltage \bar{U}_2 is greater than that of \bar{U}_1 . As a consequence, $U_4 < U_3$. The output voltage

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of the discriminator is negative in sign.

Conversely, at a lower-than-normal intermediate frequency the second circuit will have a capacitive reactance. The current \bar{I}_2 then leads the emf \bar{E} by an angle φ , as a result of which the voltage \bar{E}_2 lags behind the voltage \bar{E}_1 by $90^\circ - \varphi$ (see vector diagram in Fig.10-27c). Therefore, \bar{U}_2 is smaller than \bar{U}_1 in absolute magnitude. Correspondingly, $U_4 < U_3$, so that the discriminator output voltage becomes positive in sign.

The dependence of the output voltage of the discriminator on the intermediate

a)

b)

c)

Fig.10-27 - Discriminator Vector Diagram

a - $f_{if} > 30$ mc; b - $f_{if} = 30$ mc; c - $f_{if} < 30$ mc, which is the normal f_{if} for a receiver

frequency at its input, called the discriminator response, is shown in Fig.10-28.

At an increase in variation from midfrequency in either direction, the discriminator output voltage shows a linear increase at first; however, when detuning is very pronounced due to the resonance properties of the circuits, it again drops to zero. The discriminator response characteristic therefore displays two maximums, the linear segment between which determines the operating band for automatic tuning.

The nature of the discriminator output voltage depends on the relative magnitude of the time constants R_3C_3 and R_4C_4 , relative to the IF pulse repetition period.

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When the time constants R_3C_3 and R_4C_4 are low, the discriminator output voltage is pulseline, but when the time constants are greater it virtually retains its magnitude in the intervals between pulses, i.e., it becomes practically constant.

U_{out}

Control of Local Oscillator Frequency

As previously noted, the reflex klystron is used at the end of the decimeter band and in the centimeter band.

Fig.10-28 - Discriminator Characteristic

As it is not possible here to make a detailed check on the klystron theory in general and that of the reflex klystron in particular, let us merely cite the major physical processes occurring in the reflex klystron.

Let us imagine a constricted cavity, subject to the action of an electron stream passing through apertures (a grid) in the constricted portion (Fig.10-29). Let us assume that free HF oscillations are excited in some form within the resonator. During those fractions of a period when there is a positive charge on the grid (1) and a negative on the grid (2), the electron stream penetrating into the space between the grids is decelerated by the HF field of the resonator. The oscillating energy of the cavity is supplemented by a reduction in the energy of the decelerated electron stream. Conversely, during those portions of a period when there is a negative charge on the grid (1) and a positive on the grid (2), the electron stream in the space between the grids undergoes acceleration. The increase in the energy of the stream due to acceleration of the electrons results from a reduction in the oscillating energy of the cavity. Thus, the decelerated electron stream yields energy to the cavity, while the accelerated stream takes up energy from the cavity.

If the density of the stream is constant, the oscillating energy of the cavity

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will not be supplemented in the final analysis, since the energy of the electron stream absorbed by the cavity in the deceleration half-period, is completely returned

to the electron stream in the acceleration half-period. To have the electron stream compensate the losses in oscillating energy in the oscillator walls, the density of the electron stream must be greater during deceleration than during acceleration. In that case, the oscillations in the cavity will not be damped.

Fig.10-29 - Passage of Electron Flow Through the Grid of a Constricted Cavity, Within Which Oscillations of Constant Velocity are Sustained.

$\ominus \oplus$ represent the decelerating field of the cavity; $+ -$ represent the accelerating field thereof

Figure 10-30a gives the basic diagram of the reflex klystron. Let us assume that free oscillations are excited in the cavity by a shock, and that the energy of these oscillations has to be supplemented. The electrons passing across the alternating field between the grids of the excited

cavity in various fractions of the oscillation period and thus undergoing various changes in velocity, enter the powerful retarding field of the reflector electrode. The reflector electrode is beyond the cavity and is approximately 100 v negative with respect to the cathode. The retarding field is generated by the potential difference between cavity and reflector, which is about 300 or 400 v. Electrons entering this field lose velocity, come to a standstill, and then return to the cavity. Electrons accelerated in the field between the cavity grids penetrate closer to the electrode and therefore return to the reflector simultaneously with the electrons retarded in the field between the cavity grids; they are thus stopped and returned at a greater distance from the reflector. Consequently, the electrons returning to the cavity in a reflex klystron form clusters or bunches.

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The frequency generated may be varied, within certain limits, by changing the voltage on the reflector electrode. In actuality, if the voltage on the reflector

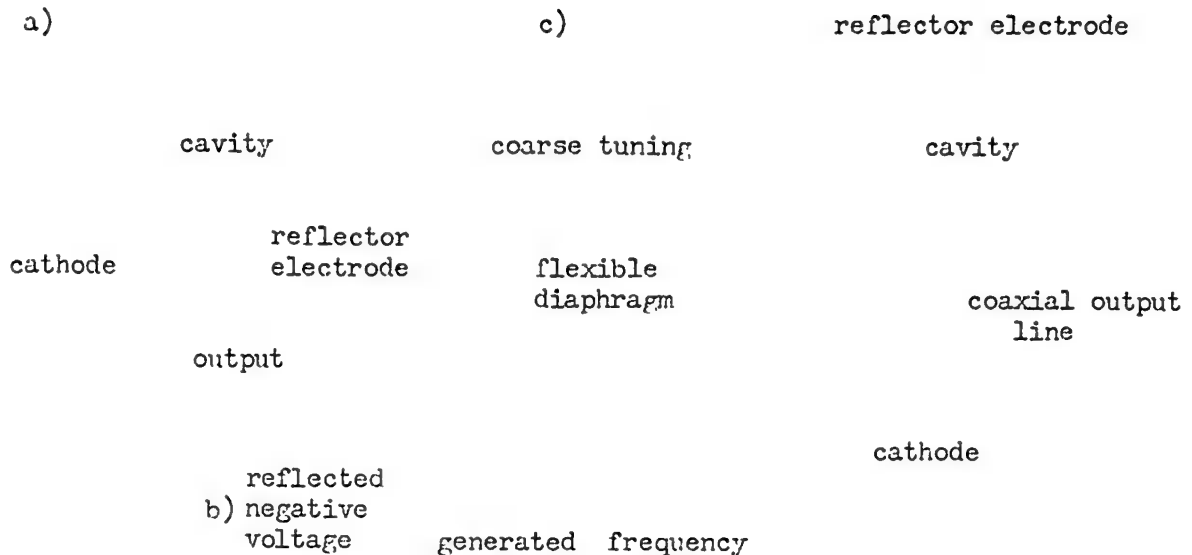


Fig.10-30 - Reflex Klystron

a - Basic diagram; b - Ratio of frequency oscillated to voltage at reflector; c - A typical design

is more negative, the bunches of electrons will return to the cavity even more rapidly. On the other hand, if the voltage on the reflector electrode is less negative, the electron bunches will return to the cavity after a longer period has elapsed. Each bunch of returning electrons transmits an energy shock to the oscillations in the cavity and compels a change in its frequency. The result is a picture similar to the change in the oscillation period of a physical lighthouse, due to external shocks not in time with its swing. An increase in the amount of negative voltage on the reflector increases the generated frequency, while a reduction in the magnitude of the negative voltage on the reflector, on the other hand, reduces the generated frequency. This results in an approximately linear relationship between the generated frequency and the voltage on the reflector. (Fig.10-30b).

Figure 10-30c presents the design of the reflex klystron.

In order to employ a klystron local oscillator in an AFC system, discrimination STAT

with high time constants are used, and frequency control is generally carried out by means of a direct-current amplifier reversing the sign of the discriminator output voltage and raising this voltage to a level adequate for control of the klystron reflector electrode. The control voltage tapped from the amplifier is added to the direct negative voltage on the reflector electrode from the feed.

The direct voltage is selected so as to provide the midfrequency selected for the receiver at the normal frequency of the transmitter. If, for some reason, the transmitter frequency has increased or the heterodyne frequency has declined, causing an increase in the intermediate frequency, the positive discriminator output voltage, amplified and reversed in sign (see Fig. 10-28) will raise the negative voltage at the reflector and increase the frequency of the local oscillator. Conversely, if the transmitter frequency has been reduced or the local oscillator frequency has increased, resulting in a reduction in the intermediate frequency, the negative output voltage of the discriminator, amplified and reversed in sign, will reduce the negative voltage on the reflector electrode and thus reduce the frequency of the local oscillator. In both cases, the normal midfrequency will be restored rather rapidly, and at the end of the compensation, the discriminator output voltage, reduced to zero, will return to the reflector the initial negative potential of the power feed.

If a triode local oscillator is involved, the control circuit includes feedback with a 90° phase shift, known as reactive feedback, the principle of which was discovered by a Soviet scientist, G.V. Braude. Control of the local oscillator frequency is obtained by means of a reactance tube, whose plate-cathode circuit is coupled parallel to the local-oscillator circuit, and serves as an aligning reactance resistance.

Figure 10-31 shows a control system along these lines. Here L_2 is the local oscillator tube, and L_1 is the control reactance tube. The capacitors C_2 and C_3 protect the tube grid circuits from direct voltage, while the choke L protects the

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feed source from high frequency. Therefore, in examining the operating principle of the circuit, these HF components will be disregarded. The oscillating potential from the local oscillator circuit $L_h C_h$ is impressed directly on the plate of the tube L_1 , and on the grid through the phase-shifting circuit $R_o C_o$. When $R_o \gg \frac{1}{\omega_g C_o}$ the current \bar{I}_o , branching into this circuit, may be taken to be in phase with the local oscillator voltage \bar{U}_a at the plate, while the voltage \bar{U}_{gf} at the grid, taken off capacitor C_o , lags behind the current \bar{I}_o and the plate voltage by 90° . In the

to the
discriminator

Fig.10-31 - Schematic Diagram for Control of the Frequency
of the Local Oscillator by Means of Reactance Tubes

pentodes, the plate voltage is controlled practically by the grid voltage alone, and is much less dependent on the plate voltage. Therefore, the plate current of the tube L_1 virtually coincides in phase with the grid voltage and lags behind the plate voltage by 90° .

In view of these phase ratios, the resistance in the plate-cathode gap has a reactive and, in the given case, an inductive character, i.e.,

$$\bar{Z}_{af} = \frac{\bar{U}_a}{\bar{I}_a} = j\omega_g L_{af} \quad (10-43)$$

The plate voltage is determined by the product of the grid voltage \bar{U}_{gf} and the transconductance S of the reactance tube. In turn, the grid voltage is the voltage

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drop from the current $I_o \approx \frac{\bar{U}_a}{R_o}$ due to resistance $\frac{1}{j\omega_g C_o}$. Taking this into consideration, eq.(10-43) readily yields the equivalent trimming inductance of the plate-cathode gap of the reactance tube:

$$L_{af} = \frac{R_o C_o}{S}$$

In automatic trimming, this inductance is regulated by changing the transconductance S of the reactance tube, under the influence of the discriminator output voltage. When the intermediate frequency is above the midfrequency, the negative potential of the discriminator shifts the quiescent point of the reactance tube into a region of lower transconductance. The trimming inductance rises, and the total inductance of the local oscillator circuit declines, while the frequency of the local oscillator increases. When the intermediate frequency is below the midfrequency, the local-oscillator frequency drops. In both cases, the midfrequency is restored.

Section 10-16. Systems of Protection from Interference

Not only echo signals from the target but all kinds of interference signals, pulsed and otherwise, reach the receiver input. Interference signals sometimes dangerously overload the receiver stages and make it impossible to observe the target signals on the indicator screen. Therefore, protection from interference is one of the most important circuiting problems in designing radar receivers.

1. Noises overloading the IF amplifier are composed of long signals of high power: reflections from local objects, clouds or, for example, powerful undamped oscillations from an interfering source. Protection from noises that overload IF amplifiers is obtained by instantaneous automatic volume control systems.

Figure 10-32 depicts one such system. When the K key is placed in the position (1), the IF amplifier (tube L_1) functions without this system. For maximum amplification of the signal from the target, without interference, the grid bias

of the tube L_1 is designed so that the initial quiescent point will fall into the area of maximum transconductance. However, if the target signal arrives simultaneously with powerful interference, then, superimposing itself on the noise, it will be in the interval of the upper bend of the characteristic curve of tube L_1 , where transconductance is zero, and will thus be cut off. In this manner, a power interference, by overloading the IF amplifier, may cause the target signal to disappear at the IF amplifier output.

An entirely different picture results if, in the presence of interference, the amplification by an IF amplifier stage is controlled by the instantaneous automatic volume control system. This is arranged by placing the key K in the position (2).

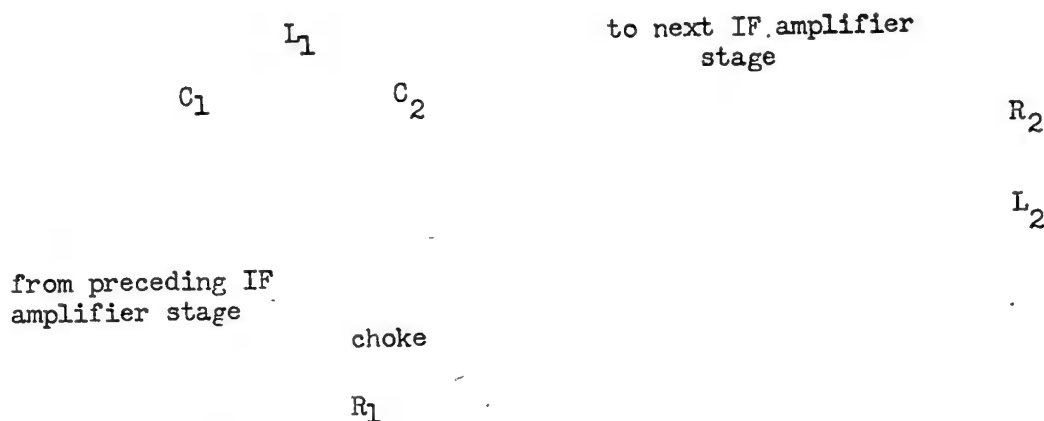


Fig.10-32 - An Instantaneous Automatic Volume Control System

The output signal of the IF amplifier stage is delivered from the plate of the tube L_1 to the grid of the tube L_2 , operated on plate detection in the lower bend of the characteristic curve. The initial grid bias required by the tube L_2 for this type of operation is provided by the divider R_4 , R_5 , and R_6 ; the grounding of the resistance R_2 with a cathode potential negative to ground is equivalent to the presence of adequate positive voltage at the plate, relative to the cathode. The grid bias of the tube L_1 , controlling the gain of the IF amplifier, is tapped

from the resistance R_2 . This is created by the plate current of the tube L_2 and is zero when that tube is off.

The increase in control bias lags somewhat relative to the instant at which the output signal of the IF amplifier appears. This lag is due to the intrinsic time $(R_1 + R_2)C_1$, which is of the same order as the transmitter pulse length. Therefore, the pure target signal is inadequate for creating a large control bias, and its amplification is not significantly reduced. However, a powerful noise of long duration, amplified by the tube L_1 , rapidly increases the control bias approximately to the level of its own amplitude, and again shifts the quiescent point of the tube L_1 into the interval of high transconductance. The target signal, superimposed on the noise after the instant when it was set by the control bias, no longer falls into the region of the upper bend of the curve, but is normally ampli-

detector

video frequency
amplifier

Fig. 10-33 - Circuit for Generation of a Short-Time Constant

fied, while the noise is suppressed to a considerable degree, thus eliminating overloading of the IF amplifier.

2. An instantaneous automatic volume control system, eliminating IF amplifier overload does not protect the video amplifier. The result of detection of noises arriving from the IF amplifier is to produce a high negative voltage on the load resistance of the detector. This may either turn off the first video amplifier altogether or cut off the target signal superimposed on the noise. The result,

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here too, is that arrival of the target signal at the receiver output again becomes impossible.

To prevent the video amplifier from being cut off by powerful interference, a low-time constant circuit ($R_g C_g$, Fig.10-33) is introduced.

When there is no prolonged interference, the key K is placed in the position(1), and the detector output is directly coupled to the video amplifier. Should interference appear, the key K is placed in the position (2), and the detector output is connected to the video amplifier via the low time-constant circuit, acting as a differential circuit. When the time constant $R_g C_g$ is somewhat smaller than the transmitter pulse length, the charging and discharging of the capacitor C_g , of brief duration, take place accordingly at the instants when the broad negative pulses of the detected noises start and end. The charging current at the beginning of interference creates a brief negative potential surge at the R_g resistance, while the discharge current at the end of interference creates a brief positive voltage surge, which is impressed on the grid of the first stage of the video amplifier. In the interval between these processes, thanks to the absence of current in the resistance R_g , the noises do not create cut-off bias at the video amplifier, and the target signal superimposed on the noise at the detector output traverses the video amplifier normally if it appears in the interval between surges.

3. A shortcoming of the low constant-time circuit lies in the fact that, although it assists the video amplifier to sustain the target signal in the presence of detected interference of long duration, it creates additional voltage surges which may confuse the operator who might interpret them as additional targets. To eliminate this drawback, systems for selecting signals by length are used. These systems pass only signals whose length is equal or nearly equal to that of the transmitter pulse. All other signals fail to pass. When the target signal is superimposed on a prolonged interference noise, only the target signal is passed, the noise being suppressed. A shortcoming of circuits for selection of signals

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by length is the suppression of grouped targets longer than the signal of a single target.

4. The power of echos from local objects and clouds near the target declines sharply with increase in range. It is therefore desirable that, in addition to connection of the instantaneous automatic volume control, there be an increase in receiver amplification with range, i.e., in proportion to lag time, so as to facilitate reception of weak signals from distant targets. A sharp reduction in gain for

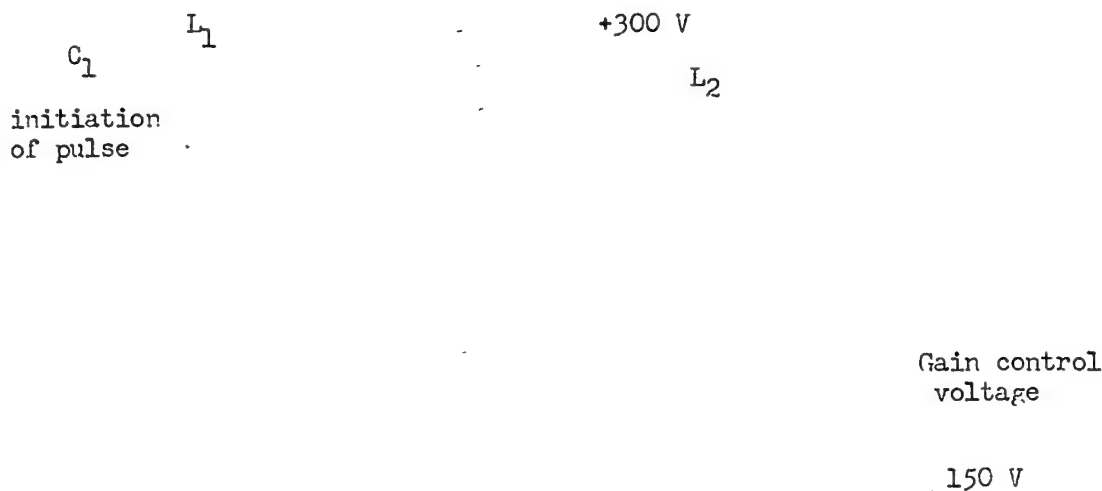


Fig.10-34 - Time-Gain Control Circuit

short ranges reduces the overload on the receiver by powerful signals near the target and assists in suppressing powerful echos from nearby local objects and clouds.

Automatic increase in gain with increasing lag time is provided by timed volume-control networks.

Figure 10-34 presents a simplified version of such a circuit. The circuit parameters are so selected that the negative plate potential of the tube L₁, taken off the divider R₆, R₈, is significantly smaller than the negative potential of its cathode taken off the divider R₃, R₄, R₅. This is equivalent to the presence of a positive voltage between plate and cathode of the tube L₁. In its initial stage,

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the tube L_1 is shut off by the high negative grid bias, while the tube L_2 remains in operation, since its cathode has a greater negative potential than the grid.

A brief positive pulse is impressed on the grid of the tube L_1 , simultaneously with the starting of the transmitter, gating it for the full duration. The plate current pulse thus set up in the tube L_1 , passing through the low-capacitance capacitor C_3 , charges it to a negative voltage whose magnitude is regulated by the potentiometer R_4 in such a fashion as not to cause the tube L_2 to be cut off.

At the instant when the starting pulse ends, tube L_1 is again ungated and the capacitor C_3 begins to discharge across the resistances R_7 and R_8 . The tube L_2 functions as a cathode follower, and the output potential on its cathode follows the negative voltage on the grid from the discharging capacitor. This is added to the negative bias taken off the potentiometer R_{10} , to control the volume. As the capacitor C_3 discharges after the triggering pulse comes to an end, the automatic negative bias on the IF amplifier declines along with the output voltage of the cathode follower, causing the receiver volume to rise with increasing range.

CHAPTER XI

GENERAL OSCILLATORS FOR RADAR TRANSMITTERS

Section 11-1. Tactical and Engineering Data on the Radar Transmitter and Specifications to be Met

The object of the radar transmitter is to generate periodic UHF current pulses of specific power, length, and shape.

Generally speaking, a radar transmitter includes three major components: oscillator, modulator, and power supply.

The oscillator portion of the transmitter consists of a UHF current oscillator, feeding the transmitting antenna. Depending on its tactical purpose and the type of radar, oscillators operate in the frequency ranges from 100 mc (3 m) to 10,000 mc (3 cm), and as a rule on fixed wavelengths.

The modulator portion of the transmitter controls its pulse operation. Pulse modulation is a specific characteristic of radar transmitters, distinguishing them from radio wavelength transmitters.

The supply source provides the transmitter with rectified voltage and feeds the filament circuit.

Due to the pulsed mode of operation, the major tactical and engineering specifications for radar transmitters are as follows:

- a) Frequency f of the oscillations to be produced (operating wavelength);
- b) Length of the R-F pulse τ_p and its shape;
- c) Train rate F_p ;

d) R-F pulse energy W_p ;

e) Transmitter pulse power P_p and its average power P_{av} during the repetition period.

The major specifications to be met by radar transmitters can be formulated as follows:

1. Provision of the required pulse power. Depending on the purpose of the station, the transmitter pulse power can range from a few watts to several megawatts. High pulse power combined with relatively low average power is a most important special feature of radar transmitters, while the inverse of the duty cycle (of the order of 1000) makes the transmitter and the radar as a whole possible, in terms of power.

2. The electric circuit of the oscillator must be as simple as possible. This requirement is dictated primarily by the difficulty and limitations of amplification of high power on ultrashort wavelengths, resulting in very low efficiency. Therefore, radar transmitter oscillators are usually single-cascade networks. Needless to say, a powerful self-exciting oscillator does not ensure the high stability of frequency provided by low-power radio broadcasting or television transmitters. The reduced frequency stability of radar transmitters requires provision of automatic tuning trim and of a pass band with an adequate safety factor, as discussed above in Section 10-6.

3. The oscillator frequency must be subject to smooth change, in tuning to a fixed operating wavelength, to provide the most desirable operating conditions for the station.

4. The R-F pulse produced must be stable in shape, amplitude, and length, and the pulse frequency must also be stable.

5. The transmitter must have a reliable control system, blocking and signaling, guaranteeing protection of operating personnel from injury by high voltage, and also ensuring simplicity of control by the transmitter and protection of its most

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important details from overloads and early wear.

6. The wiring must be of adequate mechanical strength and well protected from shock, so as to guarantee vibration stability in operation and transportation.

7. The transmitter must be light and of small dimensions, so as to render the set mobile.

Section 11-2. Basic Diagram of Radar Transmitter

In order for a powerful R-F pulse to be generated, a pulse of corresponding power must be delivered from the power supply. However, it would be exceedingly unwise to use a source with a power equal to the input pulse power. Such a source would be extremely bulky and could be used only to a very limited percentage of its capacity.

Therefore, an element for storing the energy of a low-power source is an essential component of the modulator portion of a radar transmitter. A capacitance or inductance may be used as storage element. Even if the stored energy used in the oscillator within a microsecond is negligible and does not exceed 0.6 joule, the pulse power delivered will be $P_{op} = \frac{0.6}{1 \times 10^{-6}} = 600 \text{ kw}$; if the efficiency η is 0.3, the oscillating pulse obtained will be $P_p = P_{op} = 600 \times 0.3 = 180 \text{ kw}$.

Figure 11-1 shows the basic diagram of a radar receiver.

A switch automatically couples the storage element to the power supply during the interval between pulses and to the oscillator during the time of pulse generation. The switching rate is equal to the pulse train rate.

Various methods are used for rendering the switching automatic. In the general case, the switch is controlled by the master oscillator of the set, which is not a direct component of the transmitter circuit. Although the control oscillator determines the transmitter train rate, it does not determine the shape and length of the pulse. The length and shape of the radio pulse generated depend on the rate and nature of the accumulated oscillator energy and are determined, consequently, by

the parameters of the storage element and the UHF oscillator rather than by the parameters of the control oscillator.

In certain types of stations, the control oscillator is an integral component of the transmitter. In such a case, the modulator consists of three parts: the storage element, the switch, and the control oscillator, and is used simultaneously for coordinating the functioning of all the other assemblies of the radar. The pulse modulator is often termed a keyer, as its function is that of periodic coupling and uncoupling of the oscillator.

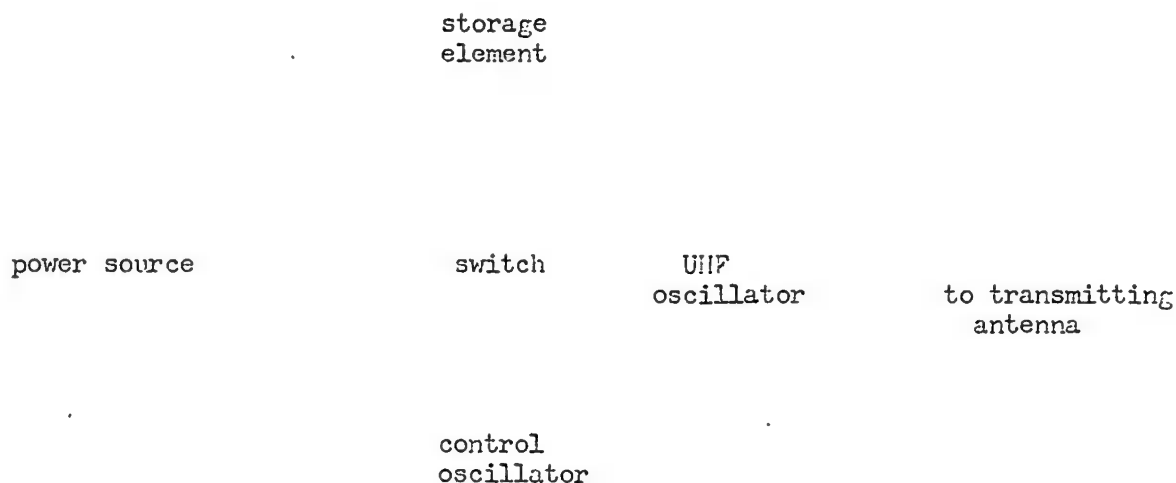


Fig.11-1 - Block Diagram of Radar Transmitter

The control oscillator proper, which is an integral component of the transmitter network, can be based on any type of oscillation or shaping of brief trigger pulses with steep leading edges (by means of a blocking oscillator, multivibrator with differentiator, limiter and amplifier, thyatron networks, etc.).

The block diagram in Fig.11-1 is generalized and is not intended to correspond to the actual circuits of the various types of radar transmitters. Sometimes, for example, it is possible to dispense with the control oscillator by coupling the storage element to the generating oscillator by means of a rotating discharger (Section 12-11), whose rotor is keyed to the shaft of a motor of a particular rpm

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of adequate stability. Finally, in certain low-power transmitters it is possible to do without the storage unit and even without the keyer at all, the oscillator tube being gated and ungated automatically in the grid circuit of the UHF oscillator itself, with parameters specifically selected for the purpose.

On the other hand, certain types of radar transmitters have block diagrams considerably more complex than those illustrated in Fig.11-1.

Section 11-3. Conditions Governing Self-Excitation of Tube Oscillator

As known from the general course on transmitters, the conditions for the self-excitation of a vacuum-tube oscillator can be reduced essentially to two:

1) The R-F voltages at the oscillator tube plate and grid must be opposite in phase. To satisfy this condition, there must be a phase balance in the components of the oscillating system of the generating oscillator;

2) The excitation voltage delivered from the plate to the grid circuit by feedback must be adequate for providing variations in plate current of a magnitude sufficient to compensate the losses in the oscillating system, when oscillations of the given frequency are generated. In order for this condition to be satisfied, the amplitudes of the alternating voltages at the plate and the grid must be balanced.

Phase Balance

The circuit of any vacuum-tube oscillator can be reduced to an equivalent tapped oscillator with self-excitation (Fig.11-2).

In accordance with the rule for tapped network designs, familiar from radio engineering, the reactance X_{gf} between grid and cathode and the reactance X_{af} between plate and cathode, must be identical in nature (both inductive or both capacitive). It is only on this condition that the current I , successively flowing around the equivalent reactance elements of the oscillating system, can keep the R-F voltages at plate and grid opposite in phase.

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Since all elements of the oscillating system are incapable of having reactions of identical nature to the oscillator frequency, the reactance X_{ag} between plate and grid must be opposite in sign to the equivalent resistance X_{ef} between grid and cathode. The total reactance of the left branch of the oscillating system has to be equal and opposite to that of the right branch. Therefore, the absolute magnitude of the reactance between plate and grid must be greater than the absolute magnitude of the reactance between grid and cathode.

Fig.11-2 - Equivalent Tapped
Network of Oscillator with
Self-Excitation

Amplitude Balance

The amplitude of the first harmonic of the plate oscillating voltage is

$$U_{ma} = I_{ma} Z_e$$

where I_{ma} is the amplitude of the first harmonic of the oscillator plate current;
 Z_e is the equivalent resistance of the oscillating system of the generator, at the oscillated frequency.

The value of U_{ma} is dependent not only on Z_e but also on the excitation voltage U_{mg} , delivered to the grid by feedback from the oscillator plate circuit. This relationship, expressed by the function $U_{ma} = F_1(U_{mg}, Z_e)$, is called the oscillating characteristic (curve 1 in Fig.11-3).

The oscillating characteristic permits determining the plate oscillating voltage, if the voltage excitation amplitude at the tube grid is known. The ascending branch of the oscillating characteristic corresponds to operation of the oscillator at less than peak voltage, when the grid current does not distort the shape of the plate-current pulse and its first harmonic increases with increasing U_{mg} . The

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descending branch corresponds to operation at excess voltage, when the appearance of considerable grid current produces a plate-current pulse dip, with the result that its first harmonic and the plate oscillating voltage decrease with increasing U_{mg} . It should be mentioned that, the greater the Z_e , the higher will be the oscillation characteristic of the oscillator.

Oscillation
characteristic

feedback curve

operation at
incomplete potential

operation at excess
potential

Fig. 11-3 - Oscillation Responses and Feedback Curves.

Regions of operation at incomplete and excess potentials
illustrated for the oscillation response 1

On the other hand, the excitation voltage U_{mg} , delivered to the grid by negative feedback, depends on the feedback factor and the oscillating voltage on the plate. Therefore, a very specific relationship exists between the oscillating voltage at the plate and the excitation voltage describing feedback. The ratio describing the function $U_{ma} = F_2(U_{mg}, k)$ is called the feedback curve.

In accordance with definition, the feedback factor is

$$k = \frac{U_{mg}}{U_{ma}}$$

from which it follows that

$$U_{ma} = \frac{1}{k} U_{mg}$$

When k is constant, the feedback curve is a straight line (line 2 in Fig. 11-3), and its angle to the U_{mg} axis determines $\tan \alpha = \frac{1}{k}$. The stronger the coupling, i.e., the greater the value of k , the smaller will be the angle of slope of the feedback curve.

The feedback curve permits determining the amplitude of the excitation voltage at a given k by means of the known oscillating voltage at the plate U_{ma} . The balance between the amplitudes U_{ma} and U_{mg} determines the point A, the intersection between the feedback curve, and the oscillation curve. Actually, it is only this point that can determine the stability of the oscillations. If the plate oscillating voltage corresponds to the point A, the excitation voltage impressed on the grid by the feedback is of the same amplitude U_{mg} at which the amplitude of the oscillating voltage at the plate corresponds to the point A. In other words, the coordinates of the point A represent a combined graphic solution of the equation for oscillation response and for feedback curve.

Thus, the conditions of oscillation of a vacuum-tube oscillator, having a given oscillation response, are determined by the feedback factor.

The feedback factor is not at all immaterial, as far as provision of oscillator self-excitation is concerned. For example, given a value of k_1 corresponding to the feedback curve 3, oscillations cannot arise in an oscillator whose oscillation response is shown as curve 1. The fact is that, regardless of the excitation voltage, the oscillation voltage at the plate in the given case excited in accordance with the oscillation curve (point B) will be smaller than the magnitude U_{ma} (point B_1) required according to the feedback curve for delivery of the initial amplitude U_{mg} to the grid. Therefore, the excitation voltage drops to a magnitude corresponding to the point V', and this in turn reduces U_{ma} to correspondence with the point V. This induces a further drop in U_{mg} and results in a rapid consecutive decrease in oscillations to zero. In order to excite the oscillator, it is necessary either to increase the feedback so as to cause its curve to slope less and

to intersect the oscillation characteristic or, without varying the feedback, to increase Z_0 to a degree that compels the oscillating response (curve 4) to pass above the feedback curve until the point of intersection is reached.

Section 11-4. The Electron Tube as UHF Oscillator

In UHF operation, the electron tube loses not only its amplifying but also its oscillating properties.

In the first place, the reduction in input impedance due to the inductive effect of the electron stream (see Section 10-5) results in an increase in the power required to be delivered by feedback from the plate to the grid circuit.

In the second place, the effect of the electron transit time is to cause the tube plate current to lag behind the grid voltage by an angle ϕ_1 . In the plate circuit, a compensating lead must be provided so as to cause the alternating voltages at the plate and grid to be in opposite phase (see point 1 of the conditions for self-excitation). It is then necessary to operate with staggered loads, and the angular argument ϕ_e of the plate-load impedance must agree with the equation $\phi_e = \phi_1$. In staggered load operation, the oscillated power separated in the plate circuit naturally decreases.

$$P_{\sim} = \frac{1}{2} U_{ma} I_{ma} \cos \phi_e \quad (11-1)$$

As the wave becomes shorter, the angular arguments ϕ_1 and ϕ_e increase. At a given frequency, which does not exceed 1000 mc for powerful oscillator triodes, ϕ_e becomes 90° . At this point, the oscillating power in the plate circuit becomes equal to zero, and the tube is no longer capable of functioning as an oscillator.

In operation with plate current cutoff, a portion of the electrons do not succeed in reaching the grid during the period of the positive half-wave in the grid voltage, and return to the cathode during the negative half-wave, subjecting it to bombardment. This significantly reduces the cathode life.

From the cathode to the grid is retarded by negative STAT

grid voltage, it yields energy to the grid circuit. On the other hand, when the electron is accelerated by positive plate voltage, on moving from grid to plate, it takes up energy from the grid circuit. The increase in electron velocity raises the dissipated power at the plate. Thus, the power required by the grid circuit is separated at the plate in the form of supplementary losses involved in heating it.

The oscillating system of a UHF generator is significantly affected by the

Fig.11-4 - Triode with Short-Circuited Plate and Grid, Used
as Oscillating System—

interelectrode capacitances of the tube and the inductance of its leads, determining the maximum oscillation frequency: the critical (wave) frequency. The critical frequency coincides with the resonant frequency of the tube, corresponding to short-circuiting of plate and grid.

It follows from Fig.11-4 that the critical frequency is

$$\omega_{\text{crit}} = \omega_0 = \frac{1}{\sqrt{(L_a + L_g) C_e}}$$

where

$$C_e = C_{ag} + \frac{C_{gf} C_{af}}{C_{gf} + C_{af}}$$

In designing a UHF oscillator tube it is necessary to provide minimum inter-electrode capacitance (to raise the critical frequency) and minimal electron

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transit time (to increase the input impedance and raise the oscillator output power).

In UHF continuous oscillation tubes, particularly those of high power, reduction in interelectrode capacitances cannot be obtained by reducing electrode surface areas, because of the need of dissipating high thermal losses. It is, therefore, necessary to proceed differently, namely to increase the distance between electrodes. In this connection, reduction in electron transit time to a minimum, when the interelectrode distances are large, creates an increase in plate voltage to tens of kilovolts, which in turn constitutes a reason why the tube vacuums must be harder.

Virtually all oscillator tubes, used in the radar band are triodes, in order to eliminate the harmful effect of inductance when screen grids and capacitances $C_{(g)g}$ are introduced between the screen and control grids, serving to increase the total capacitance of the oscillating system and to reduce the critical frequency.

Section 11-5. Special Design Features of Radar Transmitter Oscillators

A resolution of the problem of generating the shortest possible waves inevitably leads to the difficulties encountered in obtaining power sufficient for radar purposes. The critical frequency of an oscillator tube cannot be used in radar transmitters. The need for coupling to the antenna, for regulation of the oscillated frequency and determination of the most favorable operating conditions, makes it necessary to introduce the external elements of oscillating systems into electrode circuits. These elements take the form either of lumped inductances and capacitances, or of (artificial) segments of long lines, properly selected, or finally of coaxial-cylindrical cavities. Every additional reactance element added serves in all cases to reduce the oscillated frequency.

Oscillators with external artificial lines make it possible to obtain adequate power at frequencies up to 600 mc, approximately. The use of lines increases the dimensions of the oscillator. At frequencies below 150 mc, the lines required become excessively bulky, so that lumped inductances consisting of one or two coils

are used instead.

The lighthouse type of tube (see Section 11-9) is used to realize the idea of obtaining an oscillator of small size by combining oscillating circuits, tuning elements, and the oscillator tube itself into a single unit (see Section 11-9). Tubes of this type permit operation at frequencies of as high as 3000 mc (10 cm), but the oscillating power they develop provides a pulse of not over one kilowatt. Consequently, lighthouse tubes do not meet the requirements of powerful radar transmitters. For operating at wavelengths of the order of 0.5 m, tubes have been designed (see Fig. 11-12) containing oscillating lines and four pulse triodes in joint operation, providing a pulse power of up to 200 kw.

Oscillation of frequencies in excess of 1000 mc in powerful transmitters is performed exclusively by magnetrons.

Let us next discuss the problem of frequency stabilization.

The fact that the powerful oscillator circuits used in radar are single-cascade designs, prohibits usual methods of frequency stabilization by means of tourmaline or quartz plates in the master oscillator circuits and subsequent multiple frequency gain, as is done in television transmitters for example.

We know from radio engineering that the stability of an oscillator frequency is higher, the better the quality of the oscillating system. Where multicascade transmitters are concerned, this permits attainment of adequately high frequency stabilization, without the use of quartz crystals. However, in the case of radar, when a powerful self-oscillator operates directly on the antenna, the use of a high-quality oscillating system that will provide good stability is out of the question.

The low-power exciter of a multistage transmitter is intentionally unloaded so that the so-called buffer stage coupled behind it can operate without grid current. In such a case, although the efficiency of the self-oscillator is very low, this will naturally not affect the power balance of the transmitter.

The quality of the oscillating system of such an unloaded self-oscillator is

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determined only by the resistance of its own losses, and may be made high enough to provide sufficient stability.

The efficiency of a powerful radar self-oscillator, operating on an antenna, is determined to a considerable degree by the efficiency of its oscillating circuit:

$$\eta_k = \frac{r_{in}}{r + r_{in}}$$

where r is the resistance of the losses of the circuit itself;

r_{in} is the resistance introduced into the circuit from the antenna.

In order to reach high radar transmitter efficiencies efforts are being made to reduce the resistance of the losses in the circuit proper, i.e., to increase the quality of the unloaded circuit.

The quality of a transmitter circuit, normally loaded onto an antenna, is determined fundamentally by the resistance introduced, meaning it will be low. In other words, the high inherent quality of the oscillating system of a radar transmitter does not determine the stability of the frequency, an increase in which requires other measures, primarily that the transmitter be supplied with stable voltages.

On the other hand, the presence of a system of automatic alignment and of a broad reception channel, renders the question of frequency stability less pressing.

The major problems of radar transmitter oscillator design resolve, then, to the following:

- 1) Ensurance of operation at the specified wavelength with adequate pulse power;
- 2) Ensurance of maximum quality of the oscillating circuit, so as to yield the best possible efficiency;
- 3) Possibility of selecting the most desirable oscillation pattern, providing an oscillated pulse of satisfactory shape, high efficiency, required power, and adequate stability;

- 4) Possibility of smooth adjustment of the oscillated frequency within the given tolerance and without significant changes in the oscillator operating conditions.

Section 11-6. Practical Circuits for Tube Oscillators in the Meter Wavelength

In the majority of cases, vacuum-tube oscillators in the meter wavelength, operate on symmetrical output. Therefore, they are designed as a rule as single-stage push-pull networks. The discussion here will be limited to an examination of networks with artificial lines, since networks with lumped inductances were characteristic of earlier radar equipment operating on longer wave lengths and presently not of interest. In the second place, they may be equated with the same equivalent circuits of push-pull oscillators, to be examined below, as may artificial-line networks.

Oscillator with Artificial Lines in Plate, Grid, and Cathode Circuits

The network of such an oscillator (Fig. 11-5a) can be converted into a generalized equivalent circuit, where the elimination of specific elements makes it possible to obtain equivalent circuits of all subsequent vacuum tube oscillators in the meter wave band.

A segment of an end-shortened two-wire line between the tube plates has a length of somewhat less than $\frac{\lambda}{4}$ and is therefore equivalent to the coupling of an inductance between the plates. The magnitude of this inductance can be adjusted smoothly with the aid of a 'shorting bar (1). The same is true of the line segments between the grids and between the plates.

The input inductances of the artificial lines, taken together with the inductances of the tube leads, form the equivalent inductances $2L_a$, $2L_g$ and $2L_f$ of the oscillator system, coupled, respectively, between the plates, grids, and cathodes. In addition, the interelectrode capacitances C_{ag} , C_{af} , and C_{gf} enter into the oscillating system.

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If the interelectrode capacitances are considered as coupled into the external circuits of the electrodes, the equivalent circuit in Fig. 11-5b will be obtained, plotted for the first harmonic of the plate current.

a)

b)

c)

Fig. 11-5 - Oscillator with Artificial Lines in Plate,
Grid, and Cathode Circuits

a - Oscillator schematic; b - Equivalent circuit of
oscillator; c - Equivalent circuit of one arm

The tube plates are coupled to the ends of the equivalent inductance $2L_a$, with equal and opposite R-F potentials. The same holds true for the tube grids and cathodes. Consequently, the midpoints of the interelectrode inductances in the push-pull system are automatically maintained at zero potential in the R-F band.

To simplify the analysis it is convenient to subdivide the equivalent circuit of a push-pull oscillator into two completely identical arms. The equivalent circuit of the arm shown in Fig. 11-5c makes it clear that the inductances L_a , L_g , and L_c , coupled respectively to plate, grid, and cathode at the point O, are in star connection.

The theory and analysis of UHF tube generators has been developed in the works of a number of Soviet scientists, primarily by G.A. Zaytlenk and S.A. Drobov.

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The starting point of this theory is conversion of the inductance star into an equivalent inductance triangle (Fig. 11-6). In connection with the method of converting a star into a triangular connection, familiar from electrical engineering, one side of the equivalent inductance triangle is equal to a fraction, where the numerator expresses the sum of the paired products of the inductances of different rays of the star, and the denominator expresses the inductance of a ray of the star opposite to the side of the equivalent triangle in which we are interested. This rule yields a system of three formulas:

$$\left. \begin{aligned} L_{af} &= \frac{L_a L_g + L_a L_f + L_g L_f}{L_f} \\ L_{af} &= \frac{L_a L_g + L_a L_f + L_g L_f}{L_g} \\ L_{ff} &= \frac{L_a L_g + L_a L_f + L_g L_f}{L_a} \end{aligned} \right\} \quad (11-2)$$

Substitution of the star by an equivalent triangle makes it possible to convert the equivalent circuit of the arm to the form shown in Fig. 11-7, which is clearer and more convenient for purposes of analysis and design. Three important conclusions can be derived from such a circuit.

1. The oscillating system of the generating oscillator consists of three parallel circuits connected in series, and has several resonant frequencies. Therefore, oscillators of this type tend to produce several frequencies simultaneously, causing jumps from one operating frequency to another, which often interferes with the smoothness in variation of the oscillated frequency when the oscillator is retuned. At the oscillated frequency, the natural frequencies of all three circuits of the oscillating system of the oscillator are significant in effect, and depend on all the reactance parameters of the circuit.

2. Taken separately, none of the three oscillating system circuits are capable of being tuned to the frequency being generated, since otherwise the equivalent resistance would be purely active, while the oscillating system of an oscillator

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requires that the circuits be equivalent to the corresponding reactances.

3. The feedback between the grid and plate circuits of meter-wave oscillators

Fig. 11-6 - Equivalent Star (a) and Triangle b) of Inductances
in Arm Circuit

a)

b)

d)

to antenna

to antenna

pickup loop

c)

e)

to antenna

to antenna

Fig. 11-7 - Supplements to Network of Oscillator with Artificial
Lines in Plate, Grid, and Cathode Circuits

- a - Oscillating system of oscillator arm, as series connection of three oscillating circuits; b - Tapped network with inductive feedback; c - Tapped network with capacitive feedback; d - Conductive coupling to antenna; e - Inductive coupling to antenna

is complex in nature. This quantity is determined by the relationship of the

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equivalent resistances of the circuits $I_{gf}C_{gf}$ and $L_{af}C_{af}$. The equivalent inductances L_{gf} and L_{af} , entering into these circuits, corresponding to the system (11-2), depend on the ratio of inductances between electrodes of the same sign. The capacitances C_{gf} and C_{af} , strictly speaking, are determined by the corresponding inter-electrode tube capacitances and by the capacitances of the external circuit wiring, connected to the electrodes. Adjustment of feedback may have a significant effect on the operating frequency of the generator.

The conditions required for self-excitation are by no means met by any and all combinations of equivalent circuit reactances. Oscillation is possible only in a relationship of oscillating system parameters that permits operation either by means of an elementary tapped circuit with inductive feedback (Fig. 11-7b) or by means of an elementary tapped circuit with capacitive feedback (Fig. 11-7c).

Thus, regardless of the unusual design of the push-pull meter-band oscillator as compared to push-pull oscillators of long and short waves, the equivalent network of an oscillator may, in the long run, be reduced to the usual tapped system of self-excitation.

As the cathodes of oscillator tubes are DC grounded, coupling to the load (the antenna) is effected, for reasons of safety for the operating personnel, using a two-wire feeder connected to the contacts KK, which may be moved along the cathode line (Fig. 11-7d). This type of connection is equivalent to conductive coupling, the connection with the load increasing as the sliding contacts approach the cathodes.

Low-power transmitters sometimes use magnetic coupling to the antenna feeder, by means of a loop placed in the field between the plate or cathode lines (Fig. 11-7e).

The tube filament circuits of push-pull oscillators (Fig. 11-6a) also display certain peculiarities. The negligible available capacitive reactance of the capacitor C_{bt} protects the filament of the corresponding tube from further heating by UHF current. Similar protection of the filament transformer is obtained by R-F blocking chokes $L_{...}$, made for use in the meter band from spirals of silvered bare copper

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wire.

An examination of other meter-band network circuits shows that it is possible to do without special blocking chokes in the filament circuit by using the shielding properties of the cathode line within whose tube filament wires are placed. As far as the leak resistance R_f is concerned, it is usually not shunted by capacitors in radar transmitter oscillators, since the capacitance of the grid circuit wiring is adequate for the purpose.

Artificial-Line Oscillators in Plate and Grid Circuits

In the circuit of this oscillator (Fig.11-8a), as distinct from the one previously discussed, the artificial line between the cathodes is lacking. In place of the line, parasitic inductances of the cathode leads of the tube can be assumed to be cut in. Therefore, the equivalent circuit of one arm of the oscillator

a)

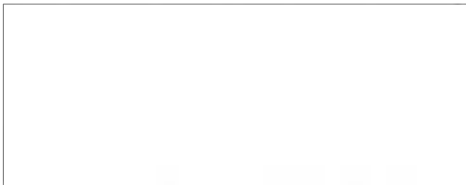
b)

c)

Fig.11-8 - Oscillator with Artificial Lines in Plate and Grid Circuits

a - Network; b - Equivalent circuit; c - Single arm circuit

may, in this case as well, be reduced to Fig.11-5c. If the inductances of the



0 cathode leads are small enough, an oscillator of the given type may operate on an equivalent tapped circuit with inductive feedback (Fig. 11-8b and c).

It is important to note that for operating by means of an inductive feedback circuit, the parallel circuits $C_{gf}L_g$ and $C_{af}L_a$ must have natural resonance frequencies higher than the oscillated frequency. Otherwise they will have no inductive reaction to the oscillated frequency. The equivalent inductance L_{af} must be larger than the equivalent inductance L_{gf} , since it governs the inductance of the right arm of the oscillating system in Fig. 11-8c. The magnitude of the inductance L_{af} basically determines the resonance frequency of the oscillating system of the oscillator. Therefore, the basic element in control of the oscillated frequency in an oscillator of this type must be considered to be the plate line. In this case, feedback is regulated by means of the grid line.

Oscillators of this type are usually employed for work in the 100-170 mc frequency band. At higher frequencies, the use of such an oscillator is not particularly desirable, since the frequency oscillated is smaller than the natural frequencies of the grid and plate circuit networks, i.e., the potentialities inherent in the oscillating system are utilized poorly in shorting the wavelength.

Oscillator with Artificial Lines in Grid and Cathode Circuits

In this type of oscillator (Fig. 11-9a), the tube anodes are connected by solid wires of the least possible length, usually in the form of tube holders firmly gripping the plate. The plates may be considered to be HF grounded to the frame across the capacitance of this holder.

Although, strictly speaking, the inductances of the plate leads are not equal to zero (the result being the circuit in Fig. 11-5b), they may be ignored for purposes of a first approximation, and it may be assumed, for simplicity in analysis, that the plates are shorted. The result is that the diagram is reduced to the equivalent push-pull network of Fig. 11-9b and to the equivalent circuit of the arm 11-9c.

Since it is only the capacitance C_{gf} that is coupled between the grid and the tube cathode in the oscillating system of the oscillator, conditions of self-excitation can exist only if the circuit $L_f C_{af}$ has only a capacitive reactance to

a)

b)

c)

d)

Fig.11-9 - Oscillator with Artificial Lines in Grid
and Cathode Circuits

a - Oscillator diagram; b - Equivalent push-pull
circuit; c - Equivalent circuit of arm; d - Hart-
self-excitation circuit

the generated frequency, while there must be only an inductive reactance in the circuit $C_{ag} L_g$. Consequently, an oscillator of this type can function only in a basic tapped system with capacitive feedback (Fig.11-9d).

It is important to note that, for operation in a capacitive feedback circuit, the parallel circuit $L_f C_{af}$ must have a resonance frequency lower than the generated frequency, since otherwise it will not have a capacitive reaction to the generated frequency. The circuit $L_f C_{af}$, on the other hand, must have a resonant frequency

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higher than that generated, to prevent an inductive response. In this connection, it need only be recalled that a parallel circuit at a frequency higher than the resonant has a capacitive response, while at lower than resonant frequency its response is inductive.

Feedback in the given circuit is determined by the ratio of the capacitance C_{gf} to the equivalent capacitance C_{afe} , and is therefore chiefly regulated, like the oscillating power, by the inductance of the cathode line. However, the length of the oscillated wave is determined primarily by the equivalent inductance L_{ag} , since the capacitance C_{gf} is invariable and the equivalent capacitance C_{afe} need only be varied within very narrow limits, in view of its strong effect on feedback and oscillation power in tuning. Thus, in an oscillator of the given type, the oscillated frequency is primarily regulated by the inductance of the grid line, and the oscillating power by the inductance of the cathode line.

As a rule, the capacitance C_{af} is lower than the capacitance C_{gf} . The feedback factor $k = \frac{C_{afe}}{C_{gf}}$ has a limiting value, $\frac{C_{af}}{C_{gf}}$ and increases with increasing L_i . On the other hand, the lower the inductance L_g , the weaker will be its shunting by the capacitance C_{ag} and the higher will be the oscillated frequency and the resultant control on any change in the inductance L_g . Therefore, L_g is always smaller than L_f , or, in other words, the grid line is always shorter than the cathode line.

The oscillator and the load (antenna) are usually coupled by means of a feeder connected to contacts that slide along the cathode line.

For R-F blocking of the filament circuits, the tube filaments in this oscillator network are shunted by the blocking capacitors C_{bH} . The cathode line tube, electrically coupled to the chassis and therefore to the grounded midpoint of the filament transformer, serves as one heater wire. The other passes within the cathode-line tube, and is therefore shielded from the high-frequency field between the cathode-line tubes. When such shielding exists, blocking chokes in the filament circuit become unnecessary.

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An oscillator with artificial lines in the grid and cathode circuits had definite advantages over one having lines in the plate and grid circuits. Primarily, within the oscillator shown in Fig. 11-9a, the cathode line is only under R-F voltage when the DC current is of zero potential. This permits direct coupling of the antenna feeder to the sliding contacts in the cathode line and facilitates smoothness in regulating the antenna coupling. Secondly, regulation of the oscillated frequency and power is here considerably simpler than in oscillators with lines in the plate and grid circuits. Thirdly, in this oscillator only the short-circuited plates which have no tuning elements, are under high direct voltage. This increases the protection from high voltage in tuning to the operating wavelength, for the operator engaged in tuning.

Section 11-7. Triode Pulse Oscillators for the Meter Band

Triodes specially designed for pulsing are used as oscillators in tube-type radar transmitters. In pulse operation, the power loss at the plate is related to the average oscillating power rather than to the pulsing. In this situation, the electrodes of the tubes, for the duration of the pulse, can withstand powers hundreds of times as great as in continuous oscillation, since they cool down during relatively long pause between pulses and heat up only during the brief period of pulse oscillation.

Therefore, electrodes in pulse triodes are rather small; this makes it possible to have low interelectrode capacitances and lead inductances. This, in turn, permits oscillation of shorter waves. Thus, triode pulse oscillators are relatively small in size, despite their high pulse powers.

The pulse power obtainable from a tube is determined by the expression

$$P_{\text{pulse}} = qE_a I_e \quad (11-3)$$

where E_a is the plate voltage;

I_e is the cathodic emission current;

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q is the tube utilization factor, usually 0.15-0.20.

In order to provide high pulse power, a tube must have the highest possible emission current, and operate at a plate voltage up to several tens of kilovolts. The maximum plate voltage for this type of tube is limited by the distance between the electrodes, the quality of the vacuum, and the inverse of the duty cycle. An increase in plate voltage is also useful for reducing the electron transit time.

Let us estimate the order of magnitude of pulse emission. For example, to provide a pulse power $P_{\text{pulse}} = 80 \text{ kw}$, the tube must, in accordance with eq.(11-3), provide

$$E_a I_e = \frac{P_{\text{pulse}}}{q} = \frac{80 \times 10^3}{0.20} = 400 \times 10^3$$

When fed with a plate voltage E_a of 8 kw, the pulse emission of the tube must be:

$$I_e = \frac{400 \times 10^3}{8 \times 10^3} = 50 \text{ amp}$$

Thus, triode pulse oscillators have to provide enormous pulse emission, ranging up to 50 amperes and more. Emission at that level is obtained by the use of carbide or oxide cathodes with enlarged surface area. Oxide cathodes are preferable since, to provide a given emission, they require less filament source power than do carbide cathodes and still less than those of tungsten.

In view of the fact that, in order to reduce interelectrode capacitances, the tube plate and grid are made as small as possible, the design of a triode pulse oscillator must provide for adequate cooling of the plate and even of the grid (usually by forced air circulation). In the operation of power tube pulse oscillators, it must be borne in mind that the presence of a powerful cathode capable of producing dangerous overheating of the adjacent grid, requires forced cooling even if the tube is only under filament voltage.

Figure 11-10 illustrates the design of a GI-17 pulse oscillator power triode.

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This tube uses a cylindrical heated oxide cathode, providing a pulse emission of $I_f = 70$ amp. The diameter of the external cathode surface is about 18 mm, measured on the oxide layer. The filament leads of the heater are made of molybdenum or tantalum and have high thermal stability. The portion of the plate within the tube is of nickel, but the external portion terminates in a large, finned surface of brass. The finned surface meets the requirement for forced air cooling. In addition, the inductances of the individual ribs are connected in parallel, making the total inductance of the plate lead equal to zero, for all practical purposes.

To reduce the effective inductance, the grid is not spiral, as is the case in long and short-wave tube oscillators, but consists of short parallel wires, connected at their ends by means of rings. The portion of the grid outside the tube

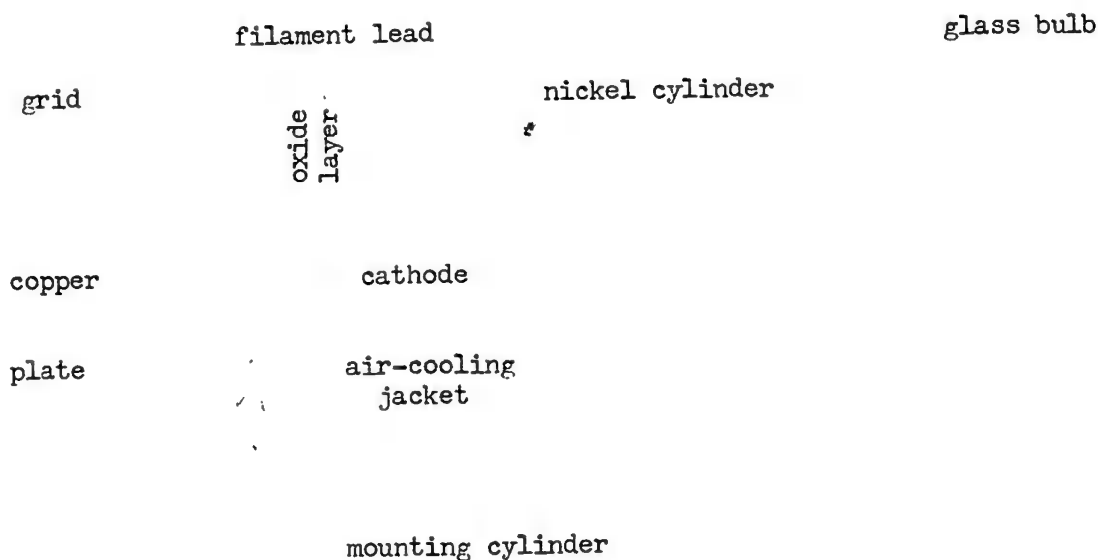


Fig.11-10 - Design of a GI-17 Triode Pulse Oscillator

terminates in a threaded copper boss to which a silvered brass terminal with air channels to cool the grid is screwed.

This design provides a 5-mm annular space between cathode and plate, the inter-

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electrode capacitances being $C_{ga} = 8 \mu\mu f$, $C_{gf} = 11 \mu\mu f$, and $C_{af} = 2 \mu\mu f$.

The electric parameters of the GI-17 triode are as follows:

$U_f = 63 \text{ v}$; $I_f = 75 \text{ amp}$; $P_f = 39 \text{ watts}$; $I_e = 70 \text{ amp}$; $E_a = 8000 \text{ v}$;

$P_{A \text{ Doppler}} = 150 \text{ watts}$; $S = 45 \text{ ma/v}$; $\mu = 15$; $\lambda_{\text{trans}} = 0.5 \text{ m}$

Section 11-8. Designing Tube Oscillators for Ultrashort Wavelengths

As distinct from long- and short-wave oscillators, the design of ultrashort wave oscillators is a product of the close coordination of the electrical and design units both as to circuiting and design. The choice of the circuit inevitably has an influence on the choice of the oscillator tubes, determining the principle followed in the engineering design of the oscillator, while the latter in turn is reflected in further refining of the circuitry.

The following considerations must be taken as the starting point in deciding on the type of tube:

- 1) Generation of the required oscillating power;
- 2) Operation at the required wavelength (satisfying the condition that $\lambda > \lambda_{\text{trans}}$);
- 3) Minimum heating power;
- 4) Desirability of selecting a tube with the lowest possible plate voltage, simplifying the high-voltage rectifier for plate feed;
- 5) Design provision for the oscillator to be employed in the proper manner;
- 6) Possibility of use in the desired method of modulation.

After the oscillator circuit and type of oscillator tube have been selected, the oscillator can be analyzed. An oscillator analysis involves two items:

- a) design of oscillator operating conditions;
- b) design of oscillator oscillating system.

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The main object of the former is to determine the equivalent resistance of the oscillating system to make sure that the given oscillating power will be obtained in the selected schedule, with satisfactory efficiency; determination of the feedback factor and determination of the resistance of the grid self-bias circuit.

The second portion of the analysis has to do with the parameters of the oscillator oscillating system, which provides tuning to the given wavelength, making it possible to obtain the feedback factor and equivalent resistance Z_e , specified under the first portion of the design, at the operating frequency.

Let us trace the major stages in the design of an ultrashort-wave push-pull oscillator by using a numerical example. Let us suppose that we have to design an oscillator on the following data:

$$P_{\text{pulse}} = 90 \text{ kw}; \tau_u = 2.5 \text{ microsec}; F_u = 400 \text{ pulses/sec}; \lambda = 1.5 \text{ m}$$

In the ultrashort-wave band, 10 to 15% of the oscillated power is lost in the oscillator tube itself. Therefore, the designed oscillating power of the oscillator must be increased accordingly to $P'_{\text{pulse}} = 1.15 \times 90 = 104 \text{ kw}$.

When an oscillator is in push-pull connection, only one arm is analyzed. Each oscillator tube must provide one-half the calculated oscillating power of the oscillator $\frac{P'_{\text{pulse}}}{2} = 52 \text{ kw}$. In the given case, we may use the GI-17 tube whose parameters have been given in Section 11-7.

Analysis of Oscillator Operating Conditions

The reasons why surge conditions must not be permitted in ultrashort-wave oscillators (despite the possibility of obtaining higher efficiency) lie not only in the pronounced reduction in useful oscillating power yielded to the oscillator when operated under conditions of overvoltage, but also in the appearance of significant grid currents which reduce the oscillator frequency stability. Therefore, it is useful to operate an ultrashort-wave oscillator at the critical level, which is characterized by maximum useful power and adequate efficiency, or under condi-

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tions of slight undervoltage close to the critical. In fact, the oscillating characteristic (Fig.11-3) shows that it is specifically at the break, corresponding to the critical schedule, at which the amplitude of swing is at its greatest.

An analysis of the operating conditions of an ultrashort-wave oscillator in terms of critical operation, or that of slight undervoltage, may be performed in the following sequence.

1. Let us select a plate voltage equivalent or close to the rated value for the given tube:

$$E_a = 8000 \text{ v}$$

2. Let us describe the utilization factor of the plate voltage $\xi = \frac{U_{ma}}{E_a}$. When a triode oscillator is operated under conditions close to the critical, we have $\xi = 0.8$ to 0.9 . Therefore, we take $\xi = 0.85$.

3. The amplitude of the oscillating voltage at the plate is

$$U_{ma} = \xi E_a$$

$$U_{ma} = 0.85 \times 8000 = 6800 \text{ v}$$

4. The phase angle of internal impedance is

$$\varphi_i^o = \varphi_e^o = \frac{360 \text{ r (mm)}}{\lambda \text{ (m)} \sqrt{E_a \text{ (v)}}}$$

where r is the distance between plate and cathode;

$$\varphi_i = \varphi_e = \frac{360 \times 5}{1.5 \sqrt{8000}} = 13.3^\circ$$

5. The amplitude of the first harmonic of the plate current is

$$I_{ma} = \frac{2 P'_{\text{pulse 1}}}{U_{ma} \cos \varphi_e}$$

$$I_{ma} = \frac{2 \times 52 \times 10^3}{6800 \times 0.973} = 15.8 \text{ amp}$$

6. Let us determine the angle of cut-off θ of the plate current. For an oscillated triode, it is desirable, from the viewpoint of good efficiency, that $\theta = 80 - 90^\circ$. We will therefore assume that $\theta = 90^\circ$.

7. Now let us use the Tables by Academician A.I.Berg to find the factor for the first plate current harmonic, $\alpha_1 = 0.5$, the factor for the constant component $\alpha_0 = 0.319$, and the factor for application of internal impedance of the tube $\alpha_i = 2$.

8. The amplitude of the plate current pulse is

$$i_{a \max} = \frac{I_{ma}}{\alpha_1}$$

$$i_{a \max} = \frac{15.8}{0.5} = 31.6 \text{ amp}$$

9. The constant component of the plate current is

$$I_{ao} = \alpha_0 i_{a \max}$$

$$I_{ao} = 0.319 \times 31.6 = 10.2 \text{ amp}$$

10. Let us calculate the critical utilization factor of the plate voltage ξ_{cr} which ensures operation in the critical conditions at a given plate current pulse:

$$\xi_{cr} = 1 = \frac{i_{a \max}}{S_{cr} E_a}$$

where S_{cr} is the transconductance of the curve for critical operating conditions and represents the geometric locus of points determining the limit of the interval for the critical schedule (Fig.11-11).

For triode oscillators, it is usual that

$$S_{cr} = 0.8 S$$

and, in our case,

$$S_{cr} = 0.8 \times 45 = 36 \text{ ma/v}$$

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Consequently,

$$\xi_{cr} = 1 - \frac{31.6}{36 \times 10^{-3} \times 8000} = 0.89$$

11. A comparison of ξ_{cr} with the value of $\xi = 0.85$, which we had assumed, shows that our oscillator operates at less than full voltage ($\xi < \xi_{cr}$), but close to the critical. Two conditions may still apply here:

- a) $\xi = \xi_{cr}$, corresponding to the critical schedule;
- b) $\xi > \xi_{cr}$, corresponding to overload conditions. To avoid overloading it would be necessary to assume a new value for ξ , somewhat smaller than ξ_{cr} , and to repeat the analysis from the start, so as to be sure of a critical or slightly overloaded set of conditions.

12. The equivalent resistance of an oscillating system at the oscillated frequency is

$$Z_e = \frac{U_{ma}}{I_{ma}}$$

$$Z_e = \frac{6800}{15.8} = 430 \text{ ohms}$$

13. The static internal impedance of the tube is:

$$R_i = \frac{\mu}{S} = \frac{15}{45 \times 10^{-3}} = 335 \text{ ohms}$$

The total internal impedance of the first harmonic of the plate current in operation at a cut-off angle of $\theta = 90^\circ$ is

$$R'_i = \alpha_i R_i$$

$$R'_i = 2 \times 335 = 670 \text{ ohms}$$

14. From the equation for the tube oscillator we find the swing of the excitation voltage:

$$U_{mg} = \frac{1}{\mu} I_{ma} (R'_i + Z_e)$$

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$$U_{mg} = \frac{1}{15} \times 15.8 (670 + 430) = 1150$$

15. The feedback factor is

$$k = \frac{U_{md}}{U_{mg}}$$

$$k = \frac{1150}{6800} = 0.16$$

Let us determine the grid self-bias voltage permitting operation at the cut-off angle selected

$$E_g = D (E_a - E_{ao}) - (U_{mg} - D U_{mg}) \cos \theta$$

when θ is 90° , we have:

$$E_g = D (E_a - E_{ao})$$

The E_a voltage, termed the reduced plate voltage, determines the plate voltage at which the idealized static characteristic $i_a = f(u_g)$ passes through the origin of coordinates. For the 6I-17 tube we have $E_{ao} \approx 500$ v, found on the basis of the tube curves. Consequently,

Fig.11-11 - Critical Schedule Curve

$$E_g = - \frac{1}{\mu} (E_a - E_{ao}) = - \frac{1}{15} (8000 - 500) = - 500 \text{ v}$$

17. The mean average oscillating power is

$$P_{\text{mean}} = \frac{P_{\text{pulse}}}{Q}$$

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$$Q = \frac{1}{\tau_u F_u} = \frac{1}{2.5 \times 10^{-6} \times 400} = 1000$$

while for a single tube we have

$$P_{\text{mean } i} = \frac{52 \times 10^3}{1000} = 52 \text{ watts}$$

18. The input power is

Pulse value

$$P_{\text{o pulse}} = I_{\text{ao}} E_a$$

$$P_{\text{o pulse}} = 10.2 \times 80000 \approx 82 \times 10^3 \text{ watts}$$

Mean value

$$P_{\text{o mean}} = \frac{P_{\text{o imp}}}{Q} = \frac{82 \times 10^3}{10^3} = 82 \text{ watts}$$

19. The mean losses per plate are

$$P_{\text{a mean}} = P_{\text{o mean}} - P_{\text{mean}}$$

$$P_{\text{a mean}} = 82 - 52 = 30 \text{ watts}$$

The permissive losses at the plate of the tube 6I-17 are 150 w, so that

$$P_{\text{a mean}} < P_{\text{a mean Doppler}}$$

20. The efficiency of the oscillator is

$$\eta = \frac{P_{\text{av}}}{P_{\text{o av}}} = \frac{52}{82} \approx 0.64$$

21. Since the circuit is push-pull and the bias voltage is created by the direct component of the grid current of two tubes, the resistance of the grid bias circuit has to be determined from the equation

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$$R_g = \frac{E_g}{2 I_{go}}$$

The magnitude of this resistance is determined by the choice of the oscillator schedule, so that there is no need for a particularly exact analysis on the basis of the grid-current curve.

For purposes of a rough estimate it may be considered that the grid current is 0.15 to 0.20 of the plate current. Therefore,

$$I_{go} = 0.15 I_{ao} = 0.15 \times 10.2 = 1.52 \text{ amp}$$

$$R_g = \frac{500}{1.52 \times 2} \approx 165 \text{ ohms}$$

For convenience in adjusting the oscillator schedule, a resistance consisting of four 50-ohm resistances connected in series, with step-by-step switching, is used.

Note: If the problem were to obtain from the oscillator the largest possible oscillating power, it would be necessary to assume the maximum plate current pulse as

$$i_{a \text{ max}} = 0.8 I_e$$

where I_e is the cathode pulse emission.

The following are then determined:

$$\xi_{cr} = 1 - \frac{0.8 I_e}{S_{cr} E_a}$$

$$U_{ma} = \xi_{cr} E_a$$

$$\left. \begin{aligned} I_{ma} &= \alpha_1 i_{a \text{ max}} \\ I_{ao} &= \alpha_o i_{a \text{ max}} \end{aligned} \right\}$$

$$P_{\text{pulse}} = \frac{1}{2} I_{ma} U_{ma} \cos \varphi_e$$

for the cut-off angle selected;

Further analysis may be carried out in the manner indicated above. At the end of this analysis, the density of the schedule is checked by comparing the magnitude of the nominal plate voltage and the maximum grid voltage

$$u_{a \min} = E_a - U_{ma}; \quad u_{g \max} = E_g + U_{mg}$$

while in the case of $u_{a \min} < u_{g \max}$, conditions of overloading exist, and if $u_{a \min} > u_{g \max}$, the situation is one of underloading. Where $u_{a \min} = u_{g \max}$, operation is in the critical schedule..

Analysis of the Oscillating System of an Oscillator

For purposes of analysis of the oscillating system of an oscillator it is necessary to draft an equivalent circuit and convert it to an elementary tapped self-excitation system. In our case, the tapped circuit of an oscillator arm is illustrated in Fig.11-10d. The distribution of the parameters of the oscillating system over its various circuits has to serve for tuning the oscillator to the required operating frequency, and also for obtaining the required Z_e and feedback factor k , determined by the foregoing analysis of the oscillator schedule.

1. In accordance with the tapped circuit (Fig.11-10d), the expression for the feedback factor

$$k = \frac{U_{mg}}{U_{ma}} = \frac{X_{gf}}{X_{afe}} = \frac{C_{afe}}{C_{gf}}$$

yields

$$C_{afe} = k C_{gf}$$

$$C_{afe} = 0.16 \times 11 = 1.76 \mu\mu f$$

2. From the equation of conductances of the circuit (Fig.11-10c)

$$\frac{1}{X_{Cafe}} = \frac{1}{X_{Caf}} - \frac{1}{X_{Lf}} \quad \text{or} \quad \frac{C_{af}}{532\lambda} - \frac{C_{afe}}{532\lambda} = \frac{1}{X_{Lf}}$$

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we find one-half the reactance component of the cathode line input impedance

$$X_{Lf} = \frac{532\lambda}{C_{af} - C_{afe}}$$

$$X_{Lf} = \frac{532 \times 1.5}{2 - 1.76} = 3300 \text{ ohms}$$

One-half the input inductance of the cathode line will be

$$L_f = \frac{X_{Lf} \lambda}{1880}$$

$$L_f = \frac{3300 \times 1.5}{1880} = 2.62 \mu h$$

3. The natural frequency of the circuit $C_{af}L_f$ is

$$f_{af} = \frac{159}{\sqrt{L_f C_{af}}}$$

$$f_{af} = \frac{159}{\sqrt{2.62 \times 2}} = 70 \text{ megacycles}$$

The resultant frequency is smaller than the oscillated frequency ($f = 200 \text{ mc}$), which is as expected.

4. The total capacitance of the oscillating system is:

$$C = \frac{C_{gf} C_{afe}}{C_{gf} + C_{afe}}$$

$$C = \frac{11 \times 1.76}{11 + 1.76} = 1.52 \mu \mu f$$

5. Ignoring the phase angle of the equivalent resistance ϕ_e , which may be corrected in tuning, we find the total inductance of the generator oscillating system, permitting operation at the given wavelength:

$$L = L_{age} = \frac{2.53 \times 10^4}{f^2 C}$$

$$L = L_{age} = \frac{2.53 \times 10^4}{200^2 \times 1.52} = 0.415 \mu h$$

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6. From the equation for conductances in the circuit

$$\frac{1}{X_{L_{oge}}} = \frac{1}{X_{L_g}} - \frac{1}{X_{C_{ag}}} \text{ or } \frac{1}{X_{L_g}} = \frac{1}{X_{L_{age}}} + \frac{1}{X_{C_{ag}}}$$

we find, first, the conductance and then one-half the reactance component of the input resistance of the grid line

$$\begin{aligned} \frac{1}{X_{L_g}} &= \frac{\lambda}{1880 L_{age}} + \frac{C_{ag}}{532 \lambda} \\ \frac{1}{X_{L_g}} &= \frac{1.5}{1880 \times 0.415} + \frac{8}{532 \times 1.5} \approx 0.012 \\ X_{L_g} &= \frac{1}{0.012} = 83 \text{ ohms} \end{aligned}$$

One-half the input impedance of the grid line is

$$\begin{aligned} L_g &= \frac{X_{L_g} \lambda}{1880} \\ L_g &= \frac{83 \times 1.5}{1880} = 0.067 \mu h \end{aligned}$$

7. The natural frequency of the circuit $L_g C_{ag}$ is

$$\begin{aligned} f_{ag} &= \frac{159}{\sqrt{L_g C_{ag}}} \\ f_{ag} &= \frac{159}{0.067 \times 8} = 220 \text{ megacycles} \end{aligned}$$

The resultant frequency is higher than that oscillated ($F = 200 \text{ mc}$), which is as should be.

After the parameters of the generator oscillating system are defined, the design analysis of the oscillating lines can be started.

As learned from practical experience, it is necessary, in line design, to strive for the optimum ratio

$$\frac{D}{d} = 4 \text{ to } 5$$

where D is the distance between the centers of the line tubes; and d is the tube diameter

diameter.

In order to ensure short-circuiting of the tube plates, and in view of the dimensions of the GI-17 tube, D has to be 60 mm. In order for the ducts not to be too thick, we set d as 12 mm. As a result, the line characteristic becomes:

$$\rho = 276 l_g \frac{2D}{d}$$

$$\rho = 276 l_g \frac{2 \times 60}{12} = 276 \text{ ohms}$$

Knowing the magnitude of the input inductance of the line L_{Bx} , the required length of the shorted portion can be determined. Considering that

$$X_{in} = \rho \tan \frac{2\pi}{\lambda} l$$

we obtain

$$l = \frac{\lambda}{2\pi} \tan^{-1} \frac{X_{in}}{\rho}$$

or

$$l \text{ (cm)} = \frac{\lambda \text{ (cm)}}{360} \left(\tan^{-1} \frac{X_{in}}{\rho} \right)^\circ$$

From our analysis of the parameters of the oscillating system, we have the following:

Input impedance of the cathode line

$$X_{in f} = 2 X_{L_f} = 2 \times 3300 = 6600 \text{ ohms}$$

Input impedance of grid lines

$$X_{in g} = 2 X_{L_g} = 2 \times 83 = 166 \text{ ohms}$$

Consequently, the length of the shorted portion of the cathode line becomes

$$l_f = \frac{150}{360} \tan^{-1} \frac{6600}{276} = 0.416 (\tan^{-1} 24)^\circ = 36.5 \text{ cm}$$

and the length of the shorted portion of the grid line will be

$$l_g = 0.416 \left(\tan^{-1} \frac{166}{276} \right)^{\circ} = 0.416 (\tan^{-1} 2.3)^{\circ} \approx 13 \text{ cm}$$

In view of the complexity of analysis of the equivalent quality of the oscillating system, it should be noted that the required equivalent resistance $Z_e = \frac{U_{ma}}{I_{ma}}$ is ensured when the oscillator tuning is controlled by the type of antenna coupling.



Fig.11-12 - Tube for $\lambda = 50$ cm, Combining Four Triodes
and One Oscillating System

Section 11-9. Microwave Oscillators

In the microwave band, the interelectrode capacitances of the tubes and the inductances of the leads constitute particularly significant components of the oscillating systems of tube oscillators. It is, therefore, highly desirable that tubes and oscillating circuits be combined into single structural elements.

Figure 11-12 gives a typical design of such a tube, providing a pulse power of about 200 kw at a wavelength of about 50 cm. In essence, such a tube is a combination of four triodes, comprising two push-pull networks operating in parallel. The oscillating system of the oscillator forms the interelectrode capacitances of these triodes and of four lines: the plate and grid, located within the tube enclosure, and two external cathode lines permitting adjustment of the oscillator schedule.

For purposes of operating at frequencies up to 3000 mc (10 cm), a tube was

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designed in 1940 by N.D.Devyatkov, Ye.N.Danil'tsev and V.K.Khokhlov, based on disk electrodes. These tubes became widely known under the name of lighthouse tubes.

The electrode leads from the glass enclosure of a lighthouse tube take the form of coaxial disks (Fig.11-13a), making it possible for lighthouse tubes to be used as removable design components of generator oscillating systems (Fig.11-13b), composed of coaxial-cylindrical cavities. The cavities are tuned by means of plungers. The disk leads are component parts of the cavities, while the plate and

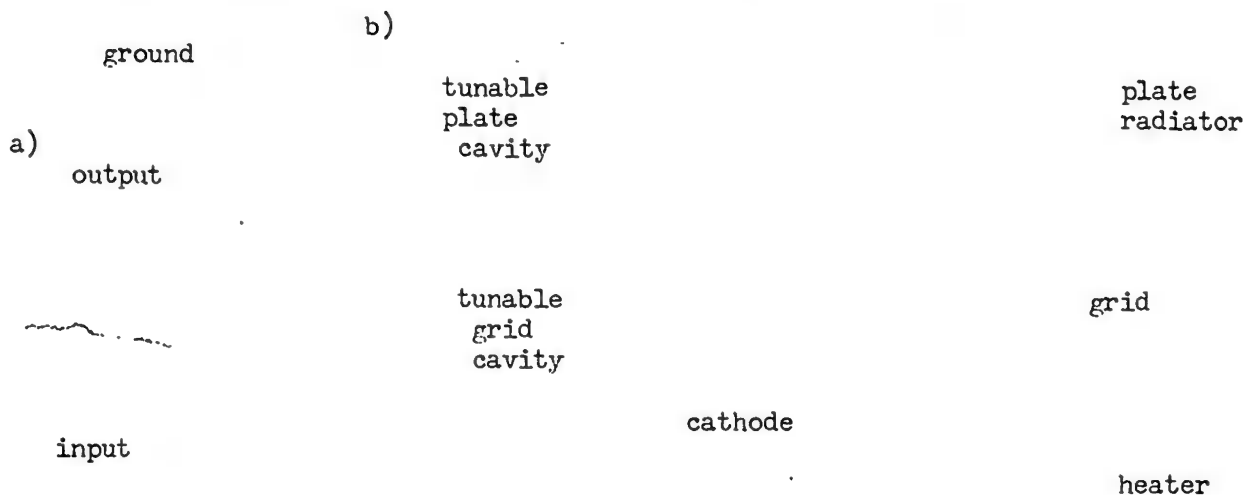


Fig.11-13 - Lighthouse Tube

Fig.11-14 - GI-7B Cermet Triode

cathode of the tube occupy part of the oscillator volume. The electrodes of lighthouse tubes are very small in surface area, so that the obtainable oscillating power does not exceed 1 kw, and the maximum permissible plate pulse voltage is of the order of 3 kw.

Using the lighthouse tubes as the starting point, a series of considerably more powerful metal and ceramic tubes was developed as early as 1944. A standard tube of this design is shown in Fig.11-14. A comparison of Figs.11-13a and 11-14 shows clearly that the cermet tube differs from N.D.Devyatkov's lighthouse tube only in that the glass has been replaced by ceramic material which is a dielectric yielding lower losses.

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Certain of the cermet tubes made in our country (the GI-7B, for example) provide a pulse power up to 20 kw at a wavelength of 11 cm. In lighthouse and cermet tubes of conventional design (Figs.11-13a and 11-14), the grid output is in a position midway between cathode and plate; therefore, the most conventional oscillator circuit for such tubes is the grounded grid.

If the internal capacitances are removed from the tube, the circuit of a self-exciting oscillator, using a grounded-grid triode, is that shown in Fig.11-15a. As it is only the C_{af} capacitance that is coupled between the tube plate and cathode, satisfaction of the conditions of self-excitation in accordance with the rule for circuiting tapped networks require that the plate-grid circuit $X_k C_{gf}$ at the oscillated frequency be equivalent to a given capacitance C_{gfe} , and the plate-grid circuit $X C_{ag}$ be equivalent to an inductance L_{af} .

Thus, the diagram of a self-exciting oscillator, using a triode with grounded

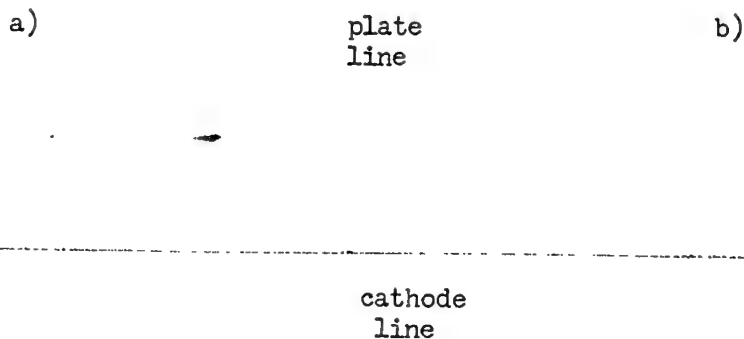


Fig.11-15 - Equivalent Circuit of Self-Exciting Oscillator

Using Grounded-Grid Triode

grid, corresponds to the equivalent tapped diagram in Fig.11-15b with capacitive feedback, the feedback factor being $k = \frac{C_{af}}{C_{gfe}}$.

Adjustment of feedback, and therefore of the oscillated power, is done by means of a plate-grid cavity.

The analysis of the schedule of a microwave oscillator is essentially analogous

to that described in Section 11-8 for the analysis of an ultrashort-wave oscillator. The only difference is that, in the case now under consideration, the two-wire lines are replaced by coaxial-cylindrical cavities, i.e., by sections of coaxial lines. Where cermet triodes are concerned, the capacitance C_{af} is 40 to 60 times smaller than the capacitance C_{ag} , so that the feedback in a circuit with grounded grid is many times weaker than in a circuit with grounded cathode.

External feedback may be performed by means of a coaxial line (Figs. 11-13b and 11-16) of adjustable length, connecting the grid and plate cavities, or by means of an adjustable supplementary capacitance between the cavities, formed by movable wire stubs containing a grid lead common to both (Fig. 11-17a) and terminating in small disks in one cavity and a wire ring in the other.

Fig. 11-16 - External Feedback by
Means of Coaxial Line of Ad-
justable Length

Sometimes, inductive feedback using special loops is employed (Fig. 11-17b), and loops being arranged symmetrically

around the perimeter of the cavity. The ends of each loop are soldered to the two sides of the surface dividing the coaxial-cylindrical cavities, and the loop wire passes through small apertures in this surface. However, this type of feedback is not adjustable, due to lack of access to the coupling loops.

Section 11-10. Magnetron Oscillators

Magnetron is the name given to an electronic device in which control of the electron stream of the cathode is effected both by the electric and magnetic fields.

Let us imagine a diode with cylindrical electrodes (Fig. 11-18), where the cavity is subject to the effects of an external magnetic field, parallel to its axis. At the same time, the lines of the electric field developed between plate and cathode

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under the influence of the plate voltage, are directed radially.

In the absence of a magnetic field, the electrons emitted by the cathode would

a)

b)

Fig.11-17 - Capacitive (a) and Inductive (b) Feedback by
Means of Special Loops

be accelerated equally along the radii (Fig.11-18a). Changes in plate current as a function of the plate voltage would be subject to Langmuir's Law, as in the ordinary diode.

magnetic
field,
H

a)

plate

cathode

b)

direction of
electron motion

lines of
magnetic force

Fig.11-18 - Diode in a Coaxial Magnetic Field

a - Orbit of electrons in the absence of a
magnetic field; b - Forces acting on the
electron

A force equal to Hev , where e is the charge on the electron, would be brought to bear by the magnetic field on an electron moving at constant velocity v in a homogeneous magnetic field H , normal to the magnetic lines of force. This force,

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normal to the direction of motion of the electrons, distorts its trajectory into a circle, equalizing the centrifugal force $\frac{mv^2}{r}$ developed as the electron moves around the circle. From the conditions for equality between these forces

$$Hev = \frac{mv^2}{r}$$

the radius of the circle described by the electron in the magnetic field can be determined, being

$$r = \frac{mv}{eh} \quad (11-4)$$

where m is the mass of the electron.

The plate voltage transmits radial accelerations to the electron. As a result

a) b) H_{cr}

of the acceleration, the radius of the electron trajectory in the magnetic field undergoes a constant increase, as it approaches the plate. At low magnetic field intensities, the appearance of the trajectory is that shown in Fig.11-19b. At sufficiently high intensities of the mag-

c) d)

netic field, the path of the electrons becomes curved to such an extent that electrons of reduced velocity do not reach the anode but, describing a closed orbit, return to the cathode. The magnetic field begins to be controlled by the plate voltage.

H_{cr} H_{cr}
Fig.11-19 - Dependence of the
Electron Orbit on the Intensity
of the Magnetic Field

At a specific magnetic field intensity, called the critical intensity H_{cr} , the plate current ceases completely. All the electrons, entering closed orbits rather epicycloid in form, return to the cathode (Fig.11-19c). The epicycloid may be

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regarded as the orbit of a point on a circle rolling on the cylindrical surface of the cathode. If, at a critical field intensity H_{cr} , the extreme points of the epicycloid continue to roll on the plate surface, they will stop rolling at $H > H_{cr}$ (Fig.11-19b).

The critical intensity of a magnetic field depends on the intensity of the electrical field, and increases therefore with increasing plate voltage and decreasing distance between plate and cathode. The greater the plate voltage, the stronger will be the electric field, while the greater the velocity of the electron, the more difficult will it be to swerve it from a linear path.

It can be shown that, where cylindrical electrons are concerned, the critical intensity of the magnetic field is

$$H_{cr} \text{ (amp/m)} = \frac{5.36}{a \text{ (cm)}} \cdot \frac{\sqrt{U_a \text{ (v)}}}{1 - \frac{b^2}{a^2}} \quad (11-5)$$

where U_a is the plate voltage, while a and b are the plate and cathode radii.

Thus, the critical intensity of the magnetic field depends on the plate voltage. When $H < H_{cr}$, the same plate voltage U_a no longer permits all electrons to return to the cathode, so that a plate current appears, whose strength U_a remaining the same, is governed by the magnitude of the intensity of the magnetic field H .

On the other hand, it is possible, at any magnetic field strength, to select a plate voltage $U_a = U_{a \text{ cr}}$ at which the given magnetic field intensity becomes critical. All electrons, having described closed epicycloids, return to the cathode, and the plate current becomes equal to zero. This plate voltage $U_{a \text{ cr}}$ is termed the critical.

Since eq.(11-5) corresponds to the condition of absence of plate current, it remains valid if the real plate voltage is taken as equal to the critical, while the critical magnetic field strength, on the other hand, is considered equal to the actual intensity H .

Substituting $U_{a \text{ cr}}$ for U_a in eq.(11-5), and H for H_{cr} , and squaring both

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parts of this equation, gives a formula for the critical plate voltage for cylindrical electrodes, as follows:

$$U_{a \text{ cr}} (v) = 3.48 \times 10^{-6} a^2 (cm) \left(1 - \frac{b^2}{a^2}\right)^2 H^2 (amp/m) \quad (11-6)$$

When $U_a > U_{a \text{ cr}}$, this same magnetic field H is no longer in a position to return all electrons to the cathode, so that of which a plate current appears, the strength of which, H being invariable, is governed by the magnitude of the plate voltage.

When the magnetic and electric fields are both effective, the strength of the plate current is a function of the magnetic field strength and of the plate voltage

$$i_a = b(H, U_a)$$

Plate-current characteristic curves expressing this dependence plotted in the absence of R-F oscillations, i.e., for static conditions, are called static magnetron characteristics or, due to their extreme steepness, clipper characteristics.

Figure 11-20a shows these characteristics in i_a, U_a coordinates, while Fig. 11-20b shows then in i_a, H coordinates.

In summarizing the foregoing, it should be emphasized that, under static conditions in the area of no plate current, all electrons emitted by the cathode describe closed epicycloid orbits and return to the cathode. Taken together, they form a space charge which takes on the nature of an electron rotor, spinning around the cathode. The velocity of rotation of this rotor, and its energy, depend on the choice of H and U_a .

Although there are a number of standard magnetron types, we will discuss only those with the multicavity resonant magnetron, the first of which in the world was developed in 1936 by the Soviet engineers N.F. Alekseyev and D.Ye. Malyarov, on the basis of an idea of that outstanding figure in radio engineering, M.A. Bonch-Bruyevich. This type of magnetron has been used exclusively in radar, as a powerful

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oscillator of very short and ultrashort waves.

The plate of a multicavity magnetron consists of a system of cavity resonators in the form of longitudinal slots (chambers) whose body has slit apertures to the

a)

b)

U_{cr1}
 U_{cr2}

H_{cr1}
 H_{cr2}

Fig.11-20 - Characteristic Clipper Curves

internal cavity (Fig.11-21a). Figure 11-21c illustrates standard types of apertures. Each resonator has distributed inductances and capacitances and serves as a cavity oscillating circuit.

In a multicavity magnetron it is primarily the inductive effect of the electron rotor that is employed. This rotor, spinning around the cathode, directs charges, varying in time, to the plate segments, and thereby excites R-F oscillations in the resonators. The final velocity at which the rotor spins is such that the oscillations in adjacent resonators do not coincide in phase. The phase shifts depend on the distance between the slots and the rate of rotation of the rotor.

The R-F fields of capacity resonators are able, due to the presence of slots which affect the DC fields created by the external magnetic system and the plate voltage, to exert a strong influence on the configuration and energy state of the electron rotor. If the electron rotor, passing a resonator, is retarded by the R-F field of the slot, it will yield part of its energy to the resonator. If on the other hand, the rotor, passing the resonator slot, is accelerated by its electromagnetic field, the rotor will absorb a portion of the resonator energy. At the

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moment of friction shown in Fig.11-21b, the slots of resonators I, III, V, and VII accelerate the motion of the rotor, and the slots II, IV, VI, and VIII retard its motion.

a)

c)

b)

rotating space
charge

Fig.11-21 - Multicavity Resonator Magnetron

a - Cutaway section; b - Electron rotor;

c - Standard types of slots

The loss of oscillating energy in each cavity is compensated by the retardation of the electron rotor which, in the final analysis, is the source of plate supply. To make it possible for the electron rotor to yield more energy to the retarding slots than it took up from the accelerating, it must approach more closely to the retarding slots than to the accelerating. The required change in the configuration of the electron rotor is reached automatically. As it approaches the accelerating slot, the electron rotor is repulsed by the negative R-F charge of the corresponding plate segment, and becomes indented. As it approaches the retarding slot, on the other hand, it picks up a further R-F charge on the corresponding segment of the plate, and is pulled toward the accelerating slot.

Thus, under dynamic conditions, i.e., in the presence of R-F oscillations, the electron rotor takes on clearly defined "peaks" facing the retarding slots,

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and "dents" opposite the accelerating slots. The electrons falling into the area of the rotor "dents" return to the cathode very rapidly. The electrons moving in the area of the "peaks" are so affected by the R-F field as to undergo deformation of the orbit from an epicycloid to a spiral; after adding to the energy of a cavity with a retarding slot, they strike the corresponding segment of the plate. Due to this fact, a certain plate current flows when the magnetron is operating under dynamic conditions and the critical intensity of the magnetic field is high, the result being that there an energy demand will be made on the plate-voltage source.

The modes operative within a magnetron are defined in accordance with the phase shift between oscillations in adjacent resonators. The most desirable are out-of-phase oscillations of the π -mode, when oscillations in adjacent resonators are mutually opposite in phase. If the rate of rotation of the electron rotor is such that the transit time between adjacent slots is one-half the oscillation period in the resonator, the rotor will undergo constant retardation, since during the transit time from the retarding slot, the formerly accelerating slot, having undergone a 180° change in phase, will also have changed to a retarding slot. When the magnetron operating conditions are properly chosen, 45 to 85% of the power required from the plate feed source is converted into R-F oscillating power. However, a portion of the oscillating power is expended on heating the resonators, so that the power yielded to the external load is only 30 to 60% of that delivered to the magnetron from the plate-voltage source.

A magnetron oscillating system consists of a large number of coupled cavity circuits, the coupling being electromagnetic in nature and carried out through the space between plate and cathode, called the interaction space. In view of the large number of coupled circuits, the oscillating system of the magnetron has a series of resonant frequencies. The choice of H and U_a governs the mode. The frequency of the magnetron is unstable during the increase in number of oscillations. In pulsed operation, should the modulated pulse have a very steep edge, the magSTATn

may skip the specified frequency and may operate at a wavelength differing greatly from that given, during the period of UHF pulse oscillation. Therefore, great care is taken in selecting the conditions of modulation of a magnetron.

In addition to the π -mode, other modes may exist in a magnetron. These are parasitic oscillations, whose presence increases the loss in the resonator chambers, reduces the efficiency of the magnetron, and diminishes the stability with which

Fig.11-22 - Straps

π -mode excitation occurs. Therefore, straps are necessary to remove the possibility of any parasitic modes being present.

With this object, magnetrons working on waves close to 10 cm length, are provided with s-called straps. These are copper rings connecting equipotential points in resonators to operate in the π -mode. The rings are attached to every other resonator (two rings at the top and two at the bottom - Fig.11-22a). In view of the equipotential state, no R-F current passes through the straps during oscillations in the π -mode. For all other modes, the points connected by the straps will no longer be equipotential. Therefore, current will flow in the straps in all other modes. The strap fields, interacting with the fields of other modes in the resonators, reduce the effective inductance of the resonators for all modes other than π . As a result, the natural frequencies of the resonators increase considerably as concerns parasitic oscillations, and excitation of parasitic oscillations is rendered more difficult, since the magnetron operating conditions (U and U_a) are based only on the conditions of excitation of frequencies similar to the main oscillation frequency mode π .

For waves close to 3 cm in length, design difficulties in the use of straps result in their replacement by a system of nonidentical pairs of resonators

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(see Fig.11-22b), to eliminate excitation of parasitic oscillations. The wavelength of the π -mode is determined by the connection of large and small resonators in parallel. The dimensions of each are so selected that the possible parasitic oscillations will be of frequencies significantly higher than the π -mode.

magnetron
resonator

coaxial
line

cavity
resonator

waveguide

to antenna

coupling loop

Fig.11-23 - Coupling of Magnetron to Antenna Line

In the π -mode, all the cavity resonators of the magnetron may be regarded as connected in parallel. The oscillating power is taken off by means of coaxial

lines coupled, when the wavelength is about 10 cm, by magnetic loops into one of the resonator cavities (Fig.11-23). A

~~cavity resonator acting as a resistance~~

transformer is used between the coaxial line and the waveguide to match the transition to the waveguide. At wavelengths of less than 3 cm, the waveguide itself is excited by means of one of the magnetron slots, with no intervening coaxial line

Fig.11-24 - Extraction of Energy from Magnetron through Slot in Waveguide (3 cm wavelength) (Fig.11-24).

The length of the stable wave, oscillated by the magnetron, is determined by the configuration and dimensions of the resonator cavities, and also by their number.

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The theory of multicavity resonator magnetrons was largely developed by the Soviet scientist, V.F. Kovalenko. It was he, also, who originated and verified in

Fig.11-25 - General View of Magnetron with Permanent Magnet

experiment the equations for resonator analysis.

The bandwidth of operating wavelengths for multicavity magnetrons of various types varies from 50 to 1 cm, although magnetrons yield less power at the shorter wavelengths, due to a reduction in the dimensions of the resonator cavities. The limiting pulse powers of magnetrons are approximately the following:

| λ | $P_{\sim \text{pulse}}$ |
|------------------|-------------------------|
| 1 cm | up to 100 kw |
| 3 cm | " " 300 kw |
| 10 cm and longer | " " 3 megawatts |

Depending on the type of magnetron, the pulsed plate voltage may vary from 1 to 45 kw. The magnetic field is provided either by a permanent magnet or by an electromagnet, the magnetron being placed between the pole pieces. The magnetic induction of the magnet is between 60 and 15,000 gauss, the decision being made in terms of the plate voltage and the operating wavelength. The plates of powerful STAT

magnetrons are water-cooled.

The magnetron cathode is usually of the oxide type. The heater heads are carefully insulated from the chassis, which is grounded. The cathode is pulsed negatively with the equivalent of the plate voltage. The operating temperature of the cathode is determined not so much by the heater as by heating of the cathode by bombardment by returning electrons.

Figure 11-25 presents a general view of a magnetron.

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